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A SPREAD SPECTRUM COMMUNICATIONS SYSTEMS UTILIZING AN UMBRELLA CODE

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THESIS

A SPREAD SPECTRUM COMMUNICATIONS SYSTEMS
UTILIZING AN UMBRELLA CODE

by

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20.
dispersive element and removed with a similar device used as a matched filter.

A Spread Spectrum Communications Systems
Utilizing an Umbrella Code

by

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ABSTRACT

A pulse stuffing technique is used to eliminate the silent gaps in the transmission of a particular Spread Spectrum system. Unauthorized listeners now have more difficulty in recovering the message content.

A two code device was built using both a pseudorandom maximal length code and a Golay complementary sequence code. The pseudo-random code was used for pulse stuffing an umbrella code and was discarded after a digital matched filter in the receiver. The Golay code was added with a Surface Acoustic Wave dispersive element and removed with a similar device used as a matched filter.

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GLOSSARY OF SYMBOLS AND TERMS

A/D	-ANALOG TO DIGITAL
BER	-BIT ERROR RATE
CCD	-CHARGE COUPLED DEVICE
CMOS	-COMPLEMENTARY METAL OXIDE SEMICONDUCTOR
CW	-CONTINUOUS WAVE
D/A	-DIGITAL TO ANALOG
IC	-INTEGRATED CIRCUIT
OOK	-ON-OFF KEYING
P/N	-PSEUDORANDOM NOISE
PRK	-PHASE REVERSAL KEYING
PCM	-PULSE CODE MODULATION
RF	-RADIO FREQUENCY
RFPA	-RADIO FREQUENCY POWER AMPLIFIER
RPU	-RECEIVE PROCESS UNIT
S/S	-SPREAD SPECTRUM
SAW	-SURFACE ACOUSTIC WAVE
SNR	-SIGNAL TO NOISE RATIO
SPU	-SEND PROCESS UNIT
TTL	-TRANSISTOR TRANSISTOR LOGIC
VCO	-VOLTAGE CONTROLLED OSCILLATOR

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I. INTRODUCTION

With the growing number of users crowding the radio spectrum, and with conventional modulation bandwidths being squeezed to fewer Hertz per channel, any scheme that promises more room ought to be investigated. Recently a new class of communications systems has developed around a modulation technique known as Spread Spectrum. Actually, spectrum spreading techniques have been around since the advent of FM. Any form of modulation added to a CW carrier will cause spreading by the addition of side bands. In these early cases, however, the bandwidth of the transmitted signal was nearly equal to that of the transmitted information. To qualify for the term "Spread Spectrum," the system must generally meet two criteria;

1) the bandwidth transmitted is usually much greater than the bandwidth of the information being sent, and

2) the transmitted bandwidth is some function other than that of the transmitted information. [1]

If the energy present in a CW signal is held constant while the signal is modulated by both information and some spreading function, the spectral power in watts per Hertz will decrease as the signal is spread wider. A point may be finally reached where the bandwidth is so wide that the spectral power is lower than the ambient noise spectral power. Within the transmission channel then, the signal is effectively buried below the noise.

The signal may be recovered by one of many coherent correlation schemes. These schemes effectively squeeze the signal in bandwidth thus bringing it above the noise again with a bandwidth equal to that of the original information. With proper amplification to recover propagation, correlation, and other losses, the original information energy can be recovered. If one is clever at designing the spreading function, it will be unique and the code period will be at least ten times the inverse of the information bandwidth. By so doing, many users may operate on the same channel and with the same bandwidth. Each user's correlation scheme will recognize and operate on only his particular code. Only the desired signal, therefore, is brought above the noise level. Other users will appear as added noise to the system (how much they add will depend on their spreading code - all other things equal). The system is not limitless. With enough users occupying the same space, a point will be reached at which the added noise overcomes the SNR required for proper operation of the system. This advantage of having many users on the same frequency and bandwidth is called "multiple access." Spread Spectrum systems offer two additional advantages, the first known as low probability of intercept or LPI, and the second known as Anti Jam or AJ. The first occurs because the signal is usually below the noise and a conventional receiver is not likely to hear it.

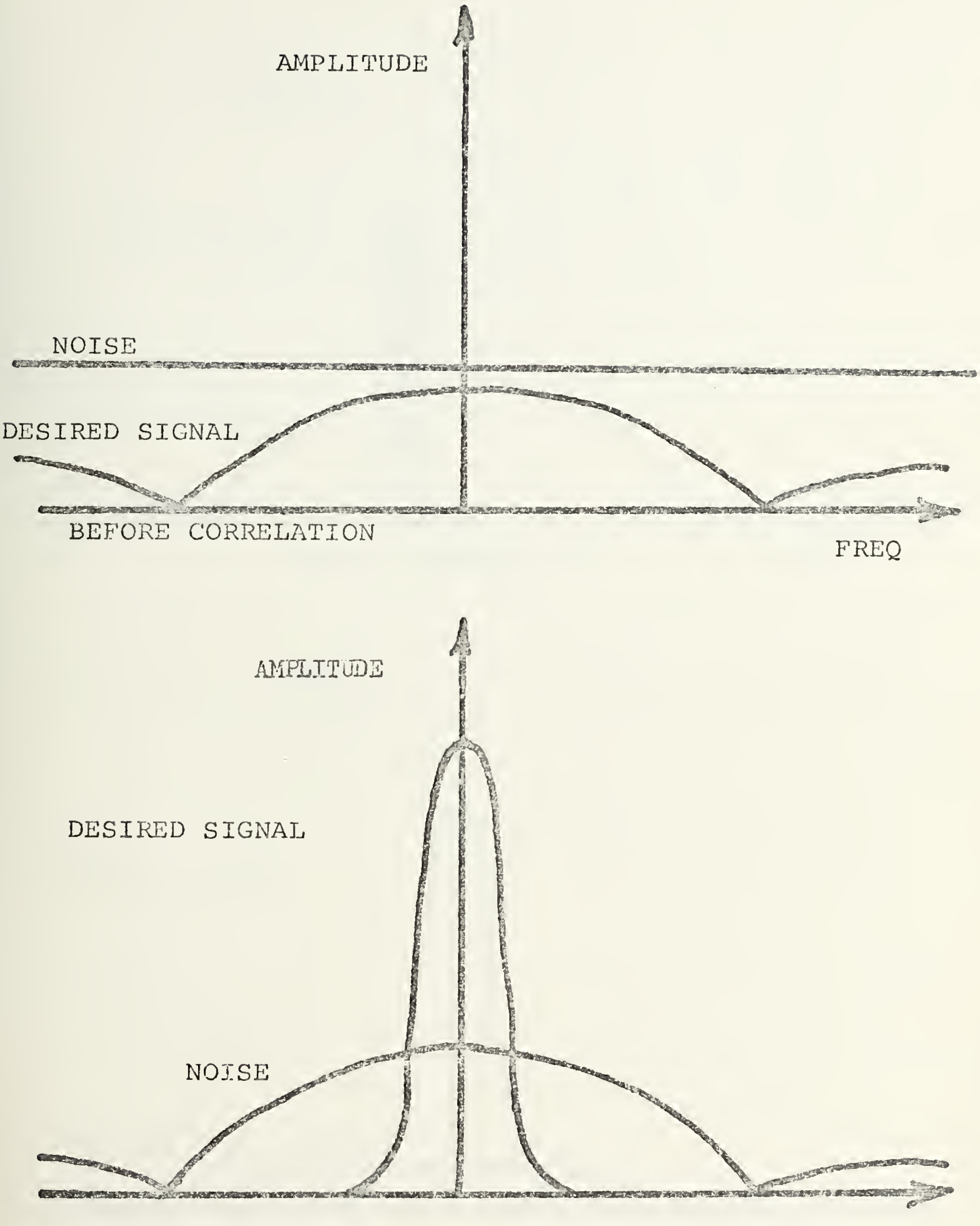
The second advaniage occurs because of the modulation process. The process used to recover the desired signal will spread interfering

signals and leave them below the level of the despread desired signal. A very strong interference is thus needed to overcome the desired signal. The level of the interfering signal is typically 20 to 30dB above that of the desired signal at the receiver before processing. Figures 1 and 2A illustrate this antijam feature.

Spread Spectrum is not without its disadvantages also. Two major disadvantages exist. The first is that a large bandwidth is required which creates difficulties in designing amplifiers and frequency converters for both the transmitter and receiver. These devices must be both broad banded and linear as most spread spectrum systems will not tolerate phase distortion.

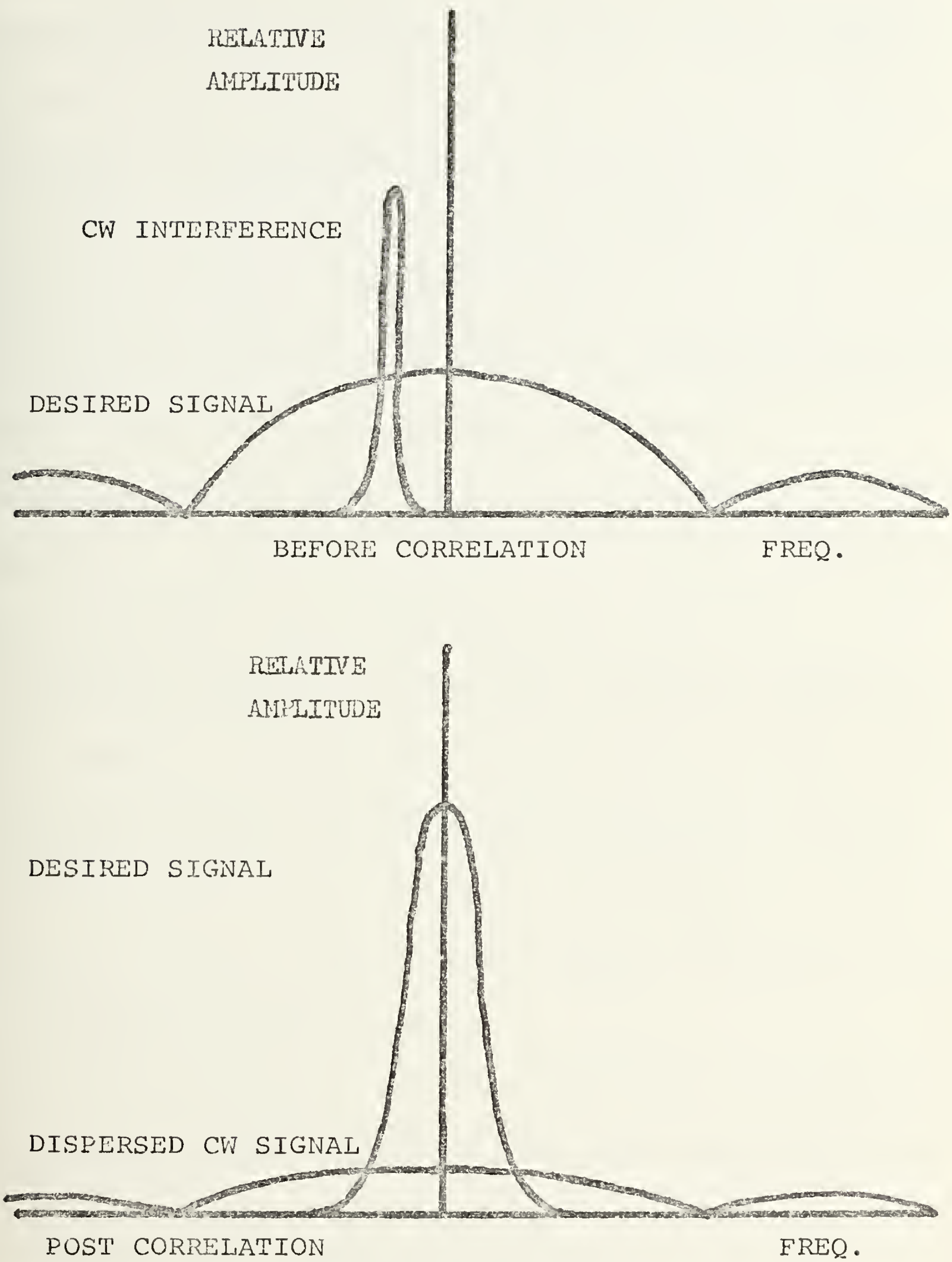
The second disadvantage is the complexity of the transmit and receive devices over that of conventional modulation systems. Complicated phase linearizing methods as well as code synchronizing schemes are necessary for proper operation.

Spread Spectrum techniques generally fall into three categories or combinations thereof. The first is chirp which is similar to the chirp techniques now used in radar applications. The second is frequency hopping in which a normally modulated carrier is transmitted for a brief time on one of a group of frequencies. The receiver must be synchronized to the hop scheme if the message is to be extracted. The third method and the one chosen for this work is direct sequence coding. Reference 2 contains a rigorous explanation of all three methods.



POWER SPECTRAL DENSITY

FIG. 1



POWER SPECTRAL DENSITY

FIG. 2 A

Less rigorously, direct sequence systems are so called because they employ a high speed (compared with information rate) spreading function code which is added in some manner to the basic information being transmitted. Usually this information is digital in form for a number of reasons such as; ease of encryption, simplification of circuitry, and a lower signal to noise ratio requirement than that for analog systems. An added advantage is that digital information is more easily handled with inexpensive IC chips which simplify circuitry over analog.

The spreading function code, being at a much faster rate than that of the information code, effectively spreads the spectrum by creating more transitions per second. Spreading function codes from a few hundred bits to megabits in length are currently used. A good rule of thumb for selecting a minimum code length to be used in a direct sequence system is to choose the sequence such that the ratio of the code bit rate to the code length is less than the lowest frequency of interest in the information being sent. In other words, the code repetition rate should not fall within the information passband. [3]

The result of modulating an RF carrier with such a spreading function is to produce a signal centered at the RF center frequency which has a

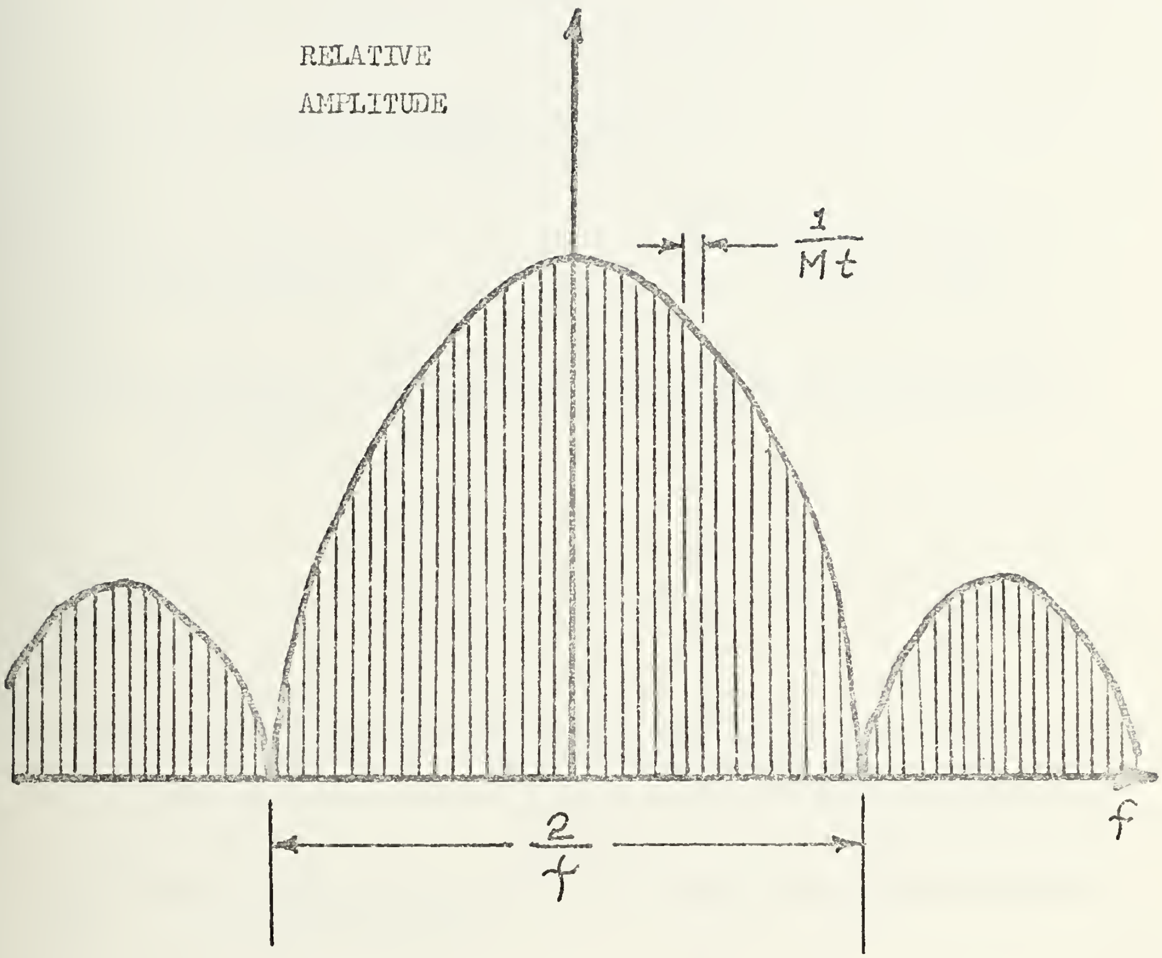
$$(\mathrm{SIN}\ X/\ X)^2$$

power spectrum. The width of the spectrum between the first nulls is twice the clock rate of the spreading function code. The side lobes will

be equal to the clock rate in width. The lines within the spectrum will occur at intervals of the code rate times the clock frequency. Figure 2B illustrates this point.

A. STATEMENT OF THE PROBLEM

Spread Spectrum systems such as designed by Cocci [4] and utilized by Garrett [5] have an inherent disadvantage of a simple spreading function and a highly visible clock rate. The entire signal would be invisible to a conventional receiver but would become apparent to an energy integrating receiver. If this receiver had a fast integration time, the clock rate might easily be determined. An easier task would be determination of the frame rate since one frame is sent per on sequence and an equally long off sequence is allowed to occur before the next frame is sent. The frame rate and, more importantly, the position of the data within the frame would be much more difficult to detect if the data were dispersed within the entire period and spaces between data bits filled with a P/N code as illustrated in Figure 3A. More than the three information bits used by Cocci could easily be sent and still have room for eight or more bits of P/N code inserted with some pulse stuffing technique. The frame length and transmit bit rate possibilities must be tempered by sample rate limitations within the A/D converter and by cost/weight limitations for a practical system.



M= THE CODE PERIOD OR TIME BETWEEN REPETITIONS

t= THE PULSE WIDTH

FREQUENCY DOMAIN REPRESENTATION OF A DIRECT SEQUENCE P/N SPREAD SPECTRUM SIGNAL. (SEE ALSO FIG. 23)

FIG. 2B

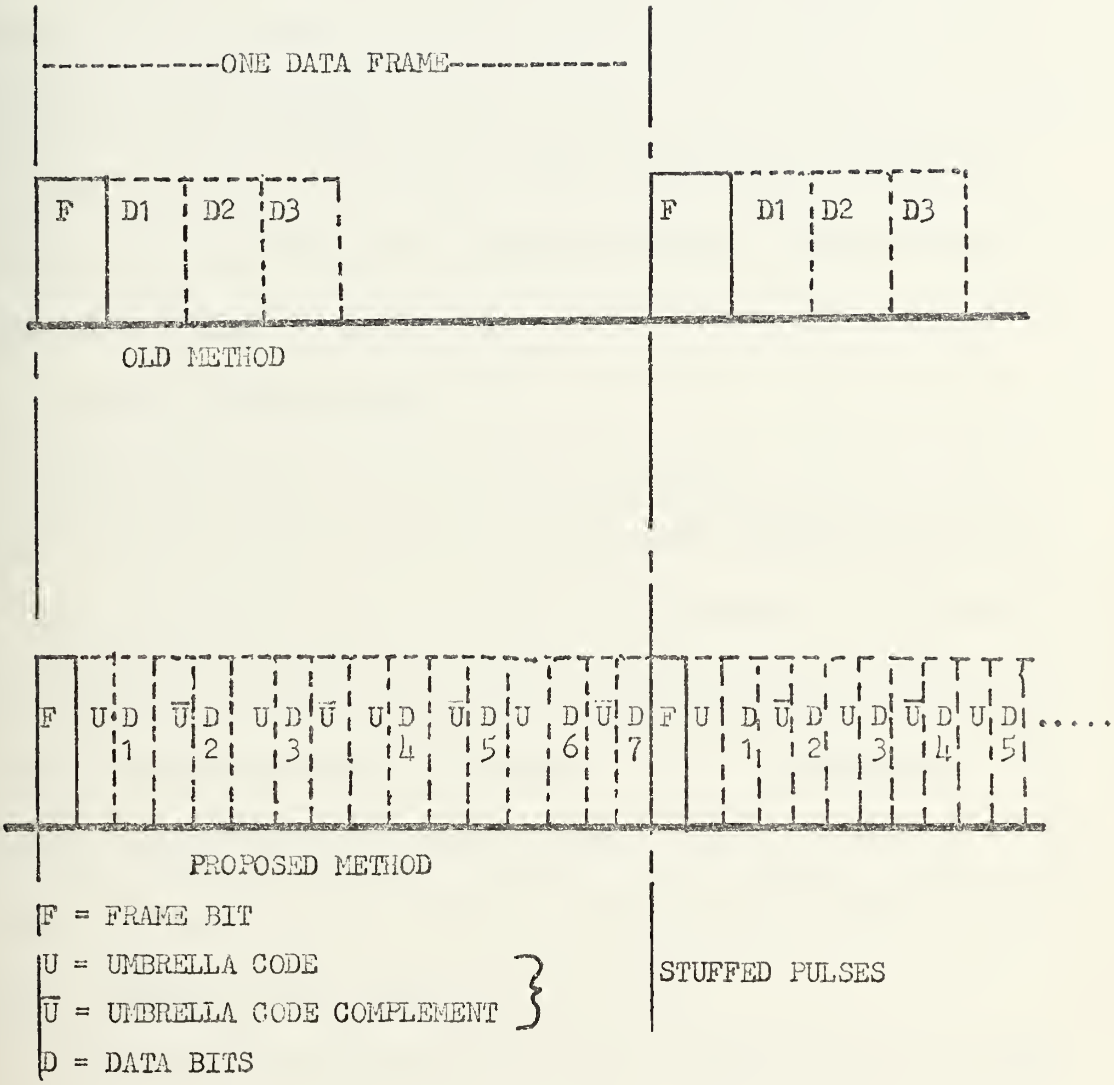

ONE DATA FRAME
F
D1
D2
D3
OLD METHOD
PROPOSED METHOD
F = FRAME BIT
U = UMBRELLA CODE
Ū = UMBRELLA CODE COMPLEMENT
D = DATA BITS
STUFFED PULSES

Fig. 3A

B. DEFINING THE THESIS

In determining how best to fill the empty spaces between information data pulses, consideration was given to modulo two addition of the spreading function code and the information code. This idea was almost immediately discarded because of the necessity of code synchronization and long delays while waiting for sync to occur. The ideal scheme would allow clock sync within two pulses and data sync within two frames. Thus, it became clear that only an unacceptably short code or a pulse stuffing scheme would achieve the desired goal.

1. General Considerations

The problem, then, was to develop a pulse stuffing technique and simultaneously restrict the size and weight of the finished product to that which would approximate a hand held transceiver. The difficulty of extracting the message content would hopefully render such a system too costly to routinely monitor on a regular basis. Additionally, with the advent of small low power crypto systems, the device should accept them with little modification. The only recourse then left for an adversary would be denial of the channel by jamming.

The entire transceiver was not constructed as it was felt that the most important part of the work was to develop the modulator-demodulator (MODEM). The MODEM should consist of a unit that would take an analog input and convert it to a digital spread spectrum signal for the transmit side. The receive side should convert the spread

spectrum digital signal to an analog signal with sufficient power to drive a speaker. The analog bandwidth should be at least 4KHz and the reproduction should have sufficient depth of modulation to reproduce both words and tonal inflection.

2. Physical Considerations

Since it was envisioned that a system of this type might eventually be incorporated into a field radio or a hand-held walkie talkie for shipboard use, major emphasis was placed on compactness, reliability, low cost and low weight. The above considerations eventually led to the choice of TTL IC chips and printed circuit boards. Some linear circuitry, also in IC form, was necessary to handle RF generation, voltage comparison and small signal amplification.

3. Theoretical Considerations

In order that the signal to noise ratio required be as low as possible, the decision was made to convert the analog signal to digital form. Since a pulse stuffing technique using fixed amplitude and fixed width of pulses had been decided on, the type of digital converter output was fixed to pulse coded modulation (PCM). An added benefit of PCM is that reconstruction of the signal at the receiver depends only on the presence or absence of a pulse rather than the pulse characteristics which are easily altered by noise. A binary state representation of 1 was chosen as a high and 0 as a low for simplicity of design.

The spreading function code schemes were somewhat more complex in design than the simple considerations given to the A/D conversion. Code synchronization should not be necessary and upon loss of signal, resynchronization should occur within the first two pulses thus allowing frame synchronization on the next frame. Noise pulses should disturb the clock sync as little as possible.

Some method of correlation was necessary in the receiver to despread the signal and force its energy above the noise level. As much processing gain as possible consistent with circuit simplicity was thus necessary. The processing gain is;

$$G_p = \frac{\text{spread Bandwidth}}{\text{information Bandwidth}} .$$

Causing the spread bandwidth to be as large as possible would maximize the processing gain.

The requirement for a low bit error rate (BER) and high noise immunity led to the use of Golay complementary coding. A device for coding an input data stream with this code was available locally in the form of a length 16 Golay coded Surface Acoustic Wave device.

Having specified the type of coding and the pulse stuffing technique to be used, the scope of the thesis was defined. The next step was the design of the necessary circuitry.

II. SPREADING FUNCTION IMPLEMENTATION

Sufficient information is presented in this section for an understanding of the MODEM design. No attempt will be made to cover surface acoustic wave devices or Golay sequences in a rigorous manner. A more complete discussion of the above topics may be found in Refs. 6, 7, 8, and 9.

A. THE BIT STUFFING TECHNIQUE

In order to intersperse the information bits with other pulses and have the resultant signal appear as random as possible, it was decided to insert the extra pulses from a pseudorandom (P/N) code generator. In making as long a code as possible, a 20 bit shift register was arranged to produce a maximal length code of 1,048,575 bits. The digital PCM information is interspersed with the P/N code and its complement, called U and $\overline{U}$, and is parallel loaded into a 16 bit parallel to serial converter shift register. (Fig. 3B) The combination is clocked out of the converter at a rate that is 16 times faster than the load rate.

At the receiver the same code comes out of the peak detector and is clocked into a 16 bit serial to parallel converter. (Fig. 3C) The pins corresponding to the U and $\overline{U}$ locations are led to exclusive OR gates, a U and a $\overline{U}$ to each gate. The first bit of the converter, which is a frame bit, and all outputs of the exclusive OR's are input to an

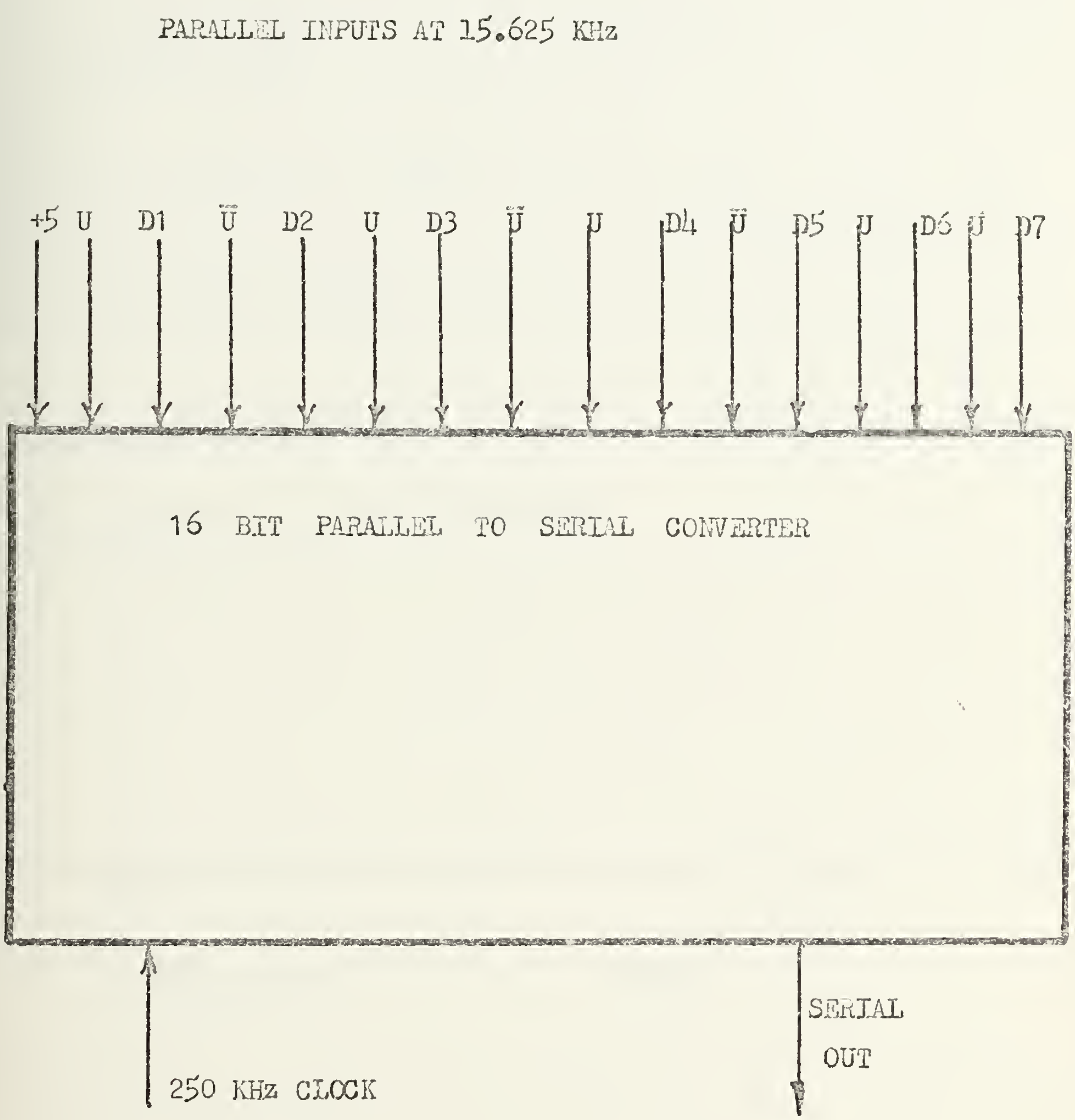

PARALLEL INPUTS AT 15.625 KHz
+5
U
D1
Ū
D2
U
D3
Ū
U
D4
Ū
D5
U
D6
Ū
D7
16 BIT PARALLEL TO SERIAL CONVERTER
250 KHz CLOCK
SERIAL OUT

Fig. 3B

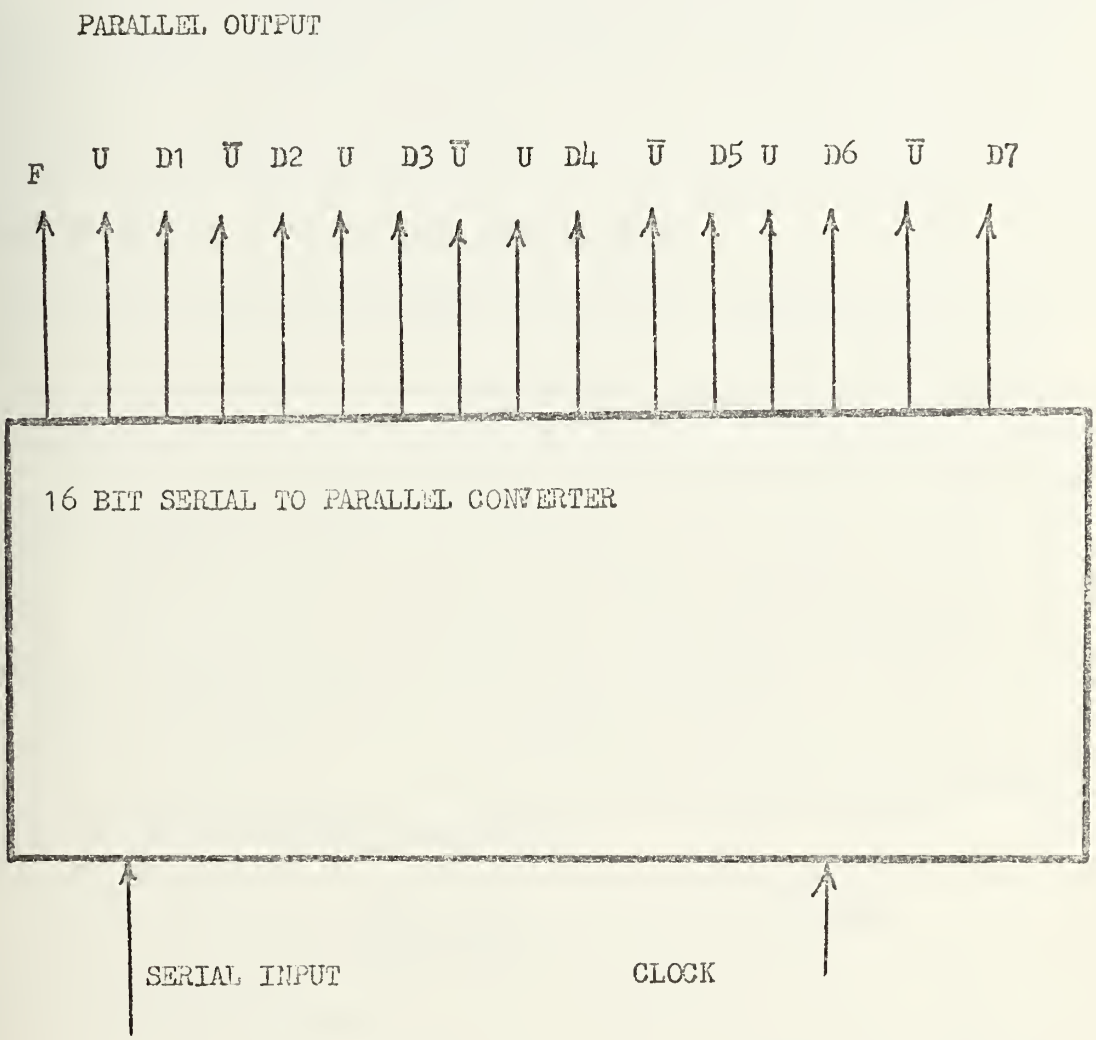

PARALLEL OUTPUT
F
U
D1
$\overline{U}$
D2
U
D3
$\overline{U}$
U
D4
$\overline{U}$
D5
U
D6
$\overline{U}$
D7
16 BIT SERIAL TO PARALLEL CONVERTER
SERIAL INPUT
CLOCK

Fig. 3C

AND gate. When all of the pulses within the data frame are in their correct position, correlation is said to occur and all inputs to the AND gate are high causing a correlation pulse output. This pulse is used as a data strobe to input the information located at the data pins of the converter to buffer memory. It then becomes a simple matter to convert the memory bits to analog form.

B. THE SURFACE ACOUSTIC WAVE DEVICE

The RF pulse generator output is connected to a Surface Acoustic Wave device which contains transducers for implementing the Golay code. This code was chosen because of its processing gain which is two times that of a direct sequence code of other types having the same bandwidth. The half power bandwidth of the Golay coded SAW device is 1.02MHz. Using the factor of two increase in Golay coded gain over that of ordinary P/N sequences, and assuming a 3KHz input information bandwidth, the processing gain would be;

$$G_p = 10 \log_{10} \frac{1.02 \times 10^6}{3.0 \times 10^3} \times 2 = 28.33\text{dB}.$$

With a direct sequence spread spectrum system, the only way the same gain could be realized would be with double the bandwidth.

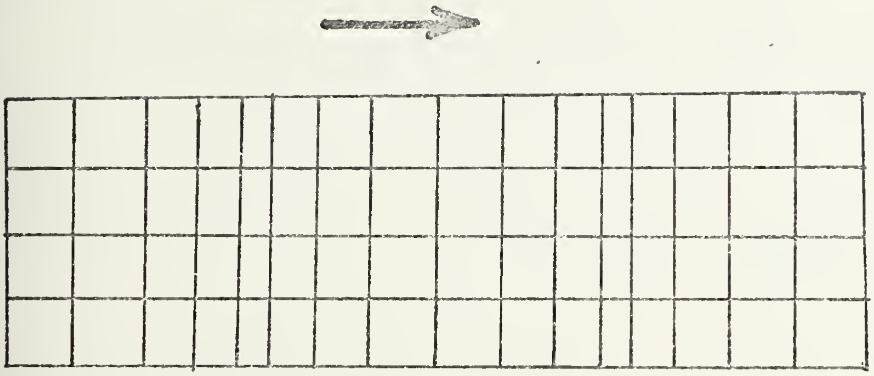
$$G_p = 10 \log_{10} \frac{2.04 \times 10^6}{3.0 \times 10^3} = 28.33\text{dB}$$

Section IV contains test results that confirm a 3dB power gain for the Golay code.

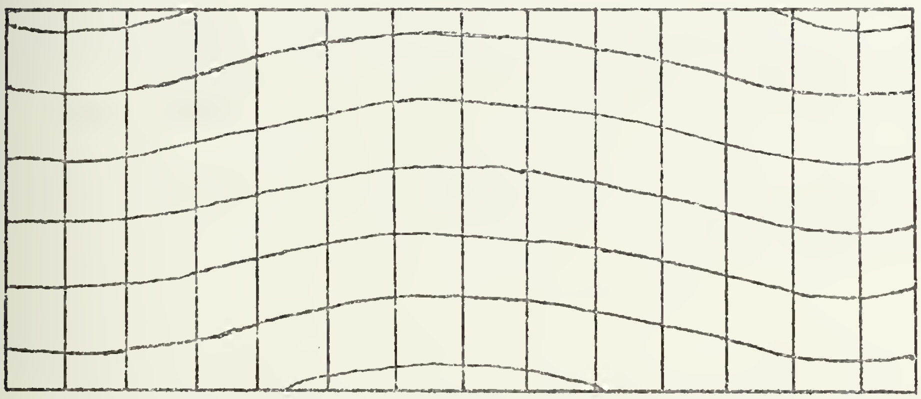
1. SAW Characteristics

In 1885, Lord Rayleigh wrote a paper predicting the existence and behavior of elastic surface waves. These waves travel on and near the surface of elastic solids with the largest component being on the surface and the amplitude exponentially decreasing with depth into the solid. The Rayleigh surface wave mode propagates in a non-dispersive manner since with most wave launchers, the aperture to wavelength ratio is large. The velocity of propagation is about five orders of magnitude less than that of light. Wave attenuation is low, being typically 1.5dB for every 10^4 wavelengths (Fig. 4). Piezoelectric solids permit the launching and detection of these Rayleigh waves by electrodes which are deposited on their surface with fabrication technology similar to that used for semiconductor devices. A real advantage of the devices is the low power that they consume, typically in the nanowatt region.

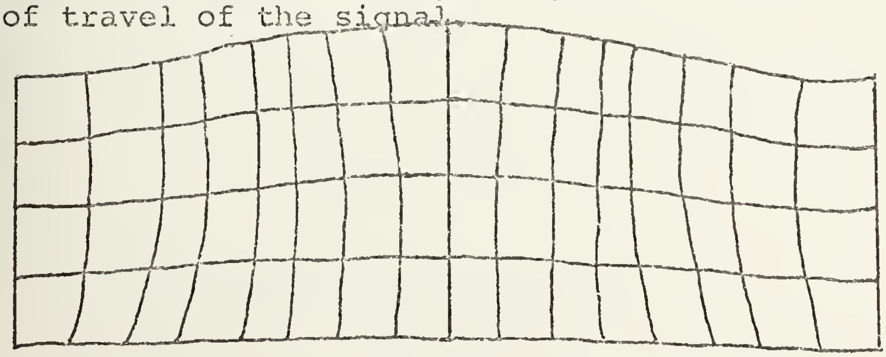
To generate a surface wave, a pulse is applied between adjacent pairs of metal fingers deposited on the surface of a piezoelectric material. The fingers are spaced a distance equal to one-half wavelength apart. (Fig. 5) The surface wave excited by the pulse travels at the surface wave velocity. By the time it reaches the next pair of launch transducer fingers, the pulse potential is caused to reverse sign. The newly excited wave will thus be in phase with the previously excited one and the amplitudes will add. After a little thought, it became obvious that the easiest manner in which a sign reversing wave could be generated was



LONGITUDINAL WAVE: the simplest kind of acoustic wave that can travel through an elastic material.



TRANSVERSE WAVE: also known as a shear wave. Oscillations are at right angles to the direction of travel of the signal.



RAYLEIGH WAVE: acoustic surface wave decays exponentially from the surface inward.

SURFACE WAVE AND OTHER MODES

FIG. 4

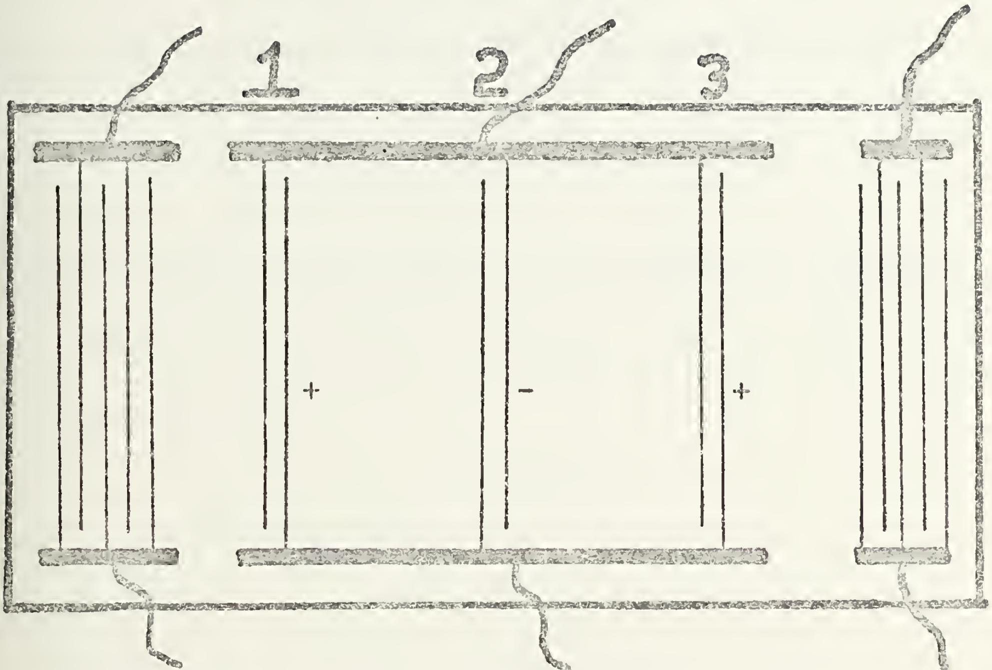
with a sine wave oscillator. This would more perfectly match the transducer since the surface waves produced were also sinusoidal. The mechanism is reciprocal and the transducer will change a mechanical wave into an electric potential. The metal finger pairs work equally well with waves arriving from either direction. Thus a finger pair set can act as either a launcher or as a receiver. (Fig. 6)

The transducer bandwidth is limited since its properties are much like the electromagnetic counterpart, the end fire antenna. A large number of finger pairs will result in a large amplitude wave out, but narrow in bandwidth since even a small change from the synchronous frequency will cause a large phase mismatch and thus small net excitation. Another bandwidth limitation may come from the matching network that couples the source to the high impedance transducer. The transducer can generally be well matched by inductively tuning out the interdigital capacitance.

Many materials can be used as the substrate in SAW devices but quartz was chosen for this one because of the weak coupling exhibited. Reflected waves and stray excitations from mechanical bumps cause less noise than if a material with better electrical coupling were used.

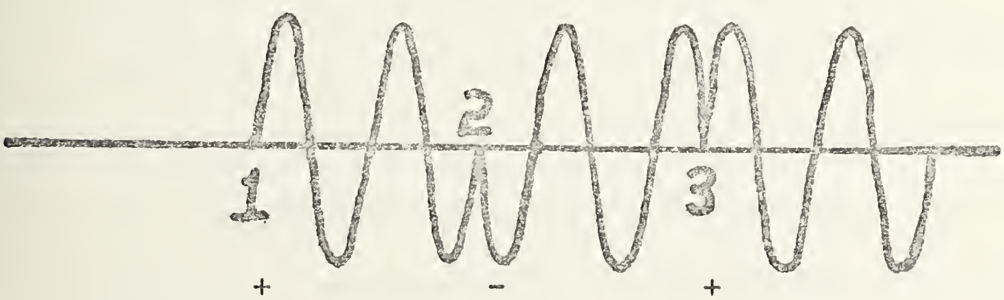
2. Golay Complementary Sequences

As defined by M. J. E. Golay, a set of complementary sequences is a pair of equally long finite sequences of two kinds of



A PHASE CODED SURFACE ACOUSTIC WAVE (SAW) DEVICE EQUIPPED AS BOTH A DISPERSER AND A MATCHED FILTER WITH LAUNCH TRANSDUCERS AT BOTH ENDS.

FIG. 5



OUTPUT OF ABOVE SAW DEVICE TAKEN AT MIDDLE LEADS. RESULT OF PHASE CODING

FIG. 6

elements which have the property that the number of like elements with any given separation between them in one sequence is equal to the number of unlike elements with the same separation in the other sequence. As an example, consider the following sequences;

A. 0 0 0 1 0 0 1 0

B. 0 0 0 1 1 1 0 1

If the separation between elements to be counted like or unlike is arbitrarily chosen as one, and if L represents like and U unlike, we have for sequence A:

0 0 0 1 0 0 1 0 = 3 likes,
L L U U L U U

and for sequence B:

0 0 0 1 1 1 0 1 = 3 unlikes.
L L U L L U U

The Golay type of complementary sequences has a useful property when applied to matched filtering. When the autocorrelation of both sequences is algebraically added, the result is all zeros except at the point of exact correlation where the sum is double that of either autocorrelation peak. The sequences are thus said to possess infinite correlation peak to peak ambiguity ratio. An example on Fig. 7 illustrates a simple Golay Complementary code and its autocorrelation. A delta function of strength eight appears at the moment of precise correlation. With a practical SAW device a delta function cannot be produced as the transducers are of finite width. The output of the SAW dispersive element will be triangular in shape which is the result of convolution of a square shaped

TIME	A*A	Raa	B*B	Rbb
1	+++- +++-	-1	++-+ ++-+	1
2	+++- +++-	-10	++-+ ++-+	10
3	+++- +++-	-101	++-+ ++-+	10-1
4	+++- +++-	-1014	++-+ ++-+	10-14
5	+++- +++-	-10141	++-+ ++-+	10-14-1
6	+++- +++-	-101410	++-+ ++-+	10-14-10
7	+++- +++-	-101410-1	++-+ ++-+	10-14-101

Raa + Rbb.......... -1 0 1 4 1 0 -1
\+ 1 0 -1 4 -1 0 1
0 0 0 8 0 0 0

AUTOCORRELATION OF A SIMPLE GOLAY COMPLEMENTARY CODE

FIG. 7

input pulse with the square SAW aperture formed by the transducer pairs. The output of the second SAW device, functioning as a matched filter, will not be a delta function but will have a second order curve shape resulting from the convolution of the input triangle with the triangular matched filter impulse response. Figures 8 and 9 illustrate the point.

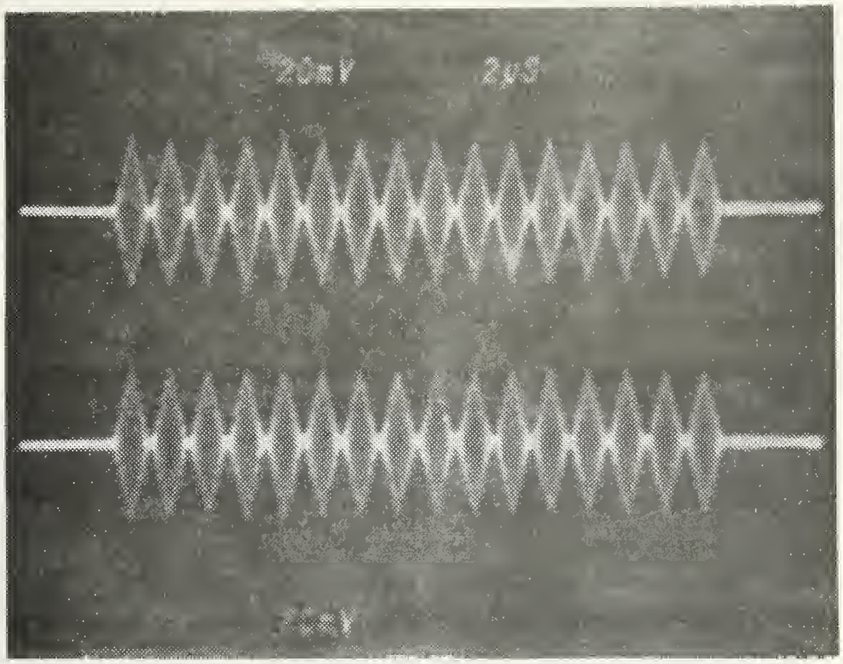
3. The Golay Sequence Used

The Golay sequences actually used in the SAW device were of length 16 and were;

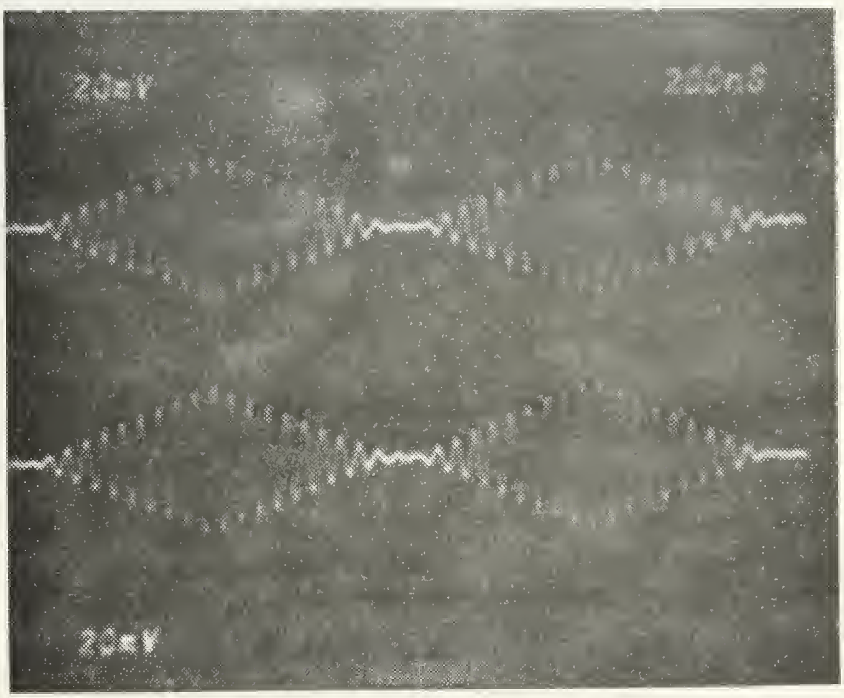
CODE A: 1 1 1 0 1 1 0 1 1 1 1 0 0 0 1 0

CODE B: 0 1 0 0 0 1 1 1 0 1 0 0 1 0 0 0

The sum of the autocorrelation functions thus had a peak of twice the code length or 32. True sidelobe cancellation was not achieved in the SAW device due to multiple reflections from the ends of the quartz substrate in spite of a graded coating of Silicone rubber applied there as an absorber. [10]



(a)



(b)

FIGURE 8 Encoder output via +20 dB.
(a) Full output, channels A & B.
(b) Last two bits, expanded.

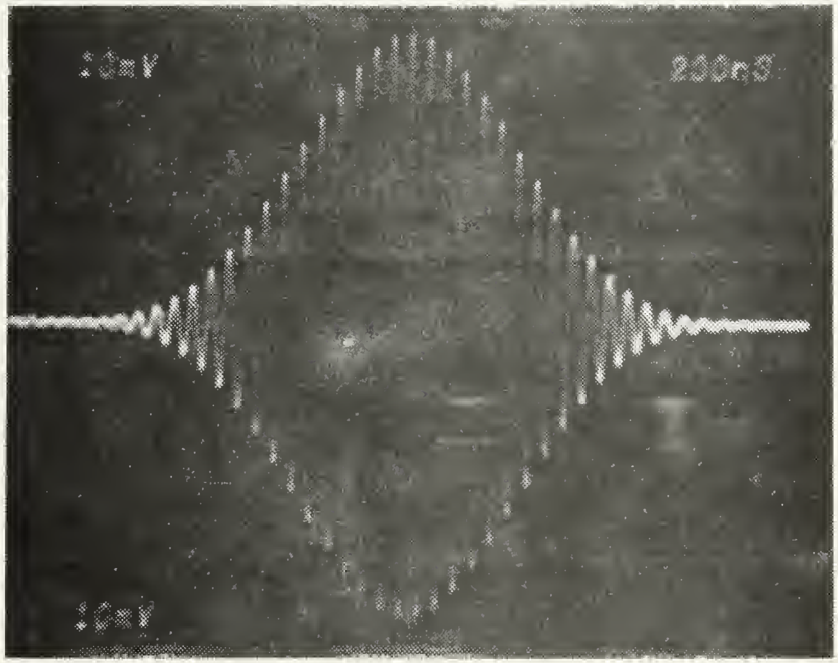


FIGURE 9 Autocorrelation sum peak via 40 dB, expanded.

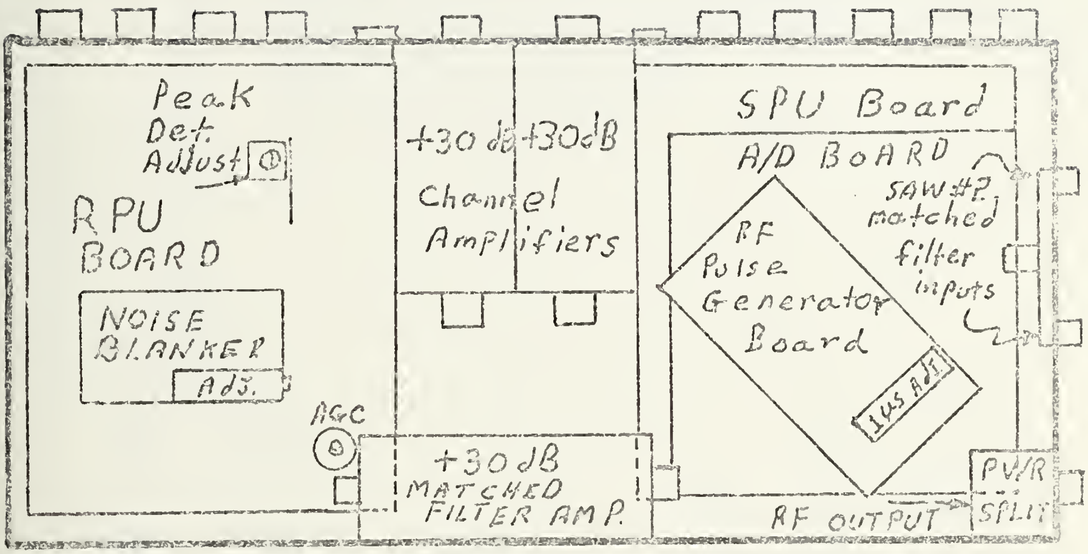
III. SYSTEM DESIGN

System design was aimed at production of a working model which would demonstrate the MODEM technique. An audio signal was to be the input. It was to be converted to digital form and then dispersed in frequency thus lowering the spectral energy. At this point additional frequency up conversion and wide band RF power amplification could be added and the signal transmitted. The received signal was to be applied to a SAW matched filter after amplification. The correlation sum was to be input to some form of peak detector and then stripped of the unwanted stuffed pulses, converted to analog and amplified. The entire device, with the exception of the SAW devices, was to be built in a box slightly larger than that of a hand held transceiver. For ease of testing, many test points were brought out with RF connectors. Figure 10 depicts the actual size and printed circuit board layout of the finished product.

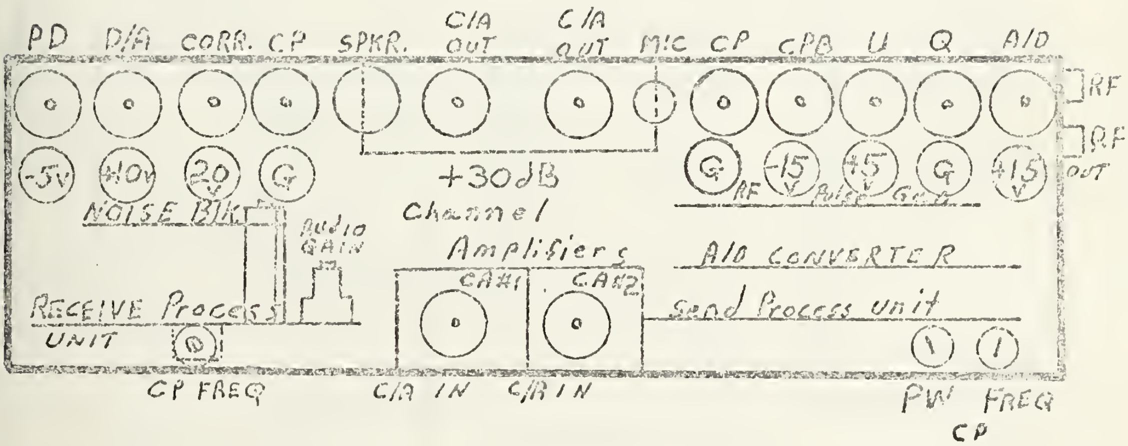
A. BLOCK DIAGRAM CONCEPT

Figure 11 shows the send process unit (SPU) block diagram. An audio signal from a crystal microphone is amplified and fed to an analog to digital converter which samples the analog signal and converts it to a PCM output. The A/D converter holds the sample until the next convert start pulse arrives thus eliminating the need for buffer storage.

TOP VIEW



SIDE VIEW



COMPLETED DEVICE COMPONENT LOCATION

FIG. 10

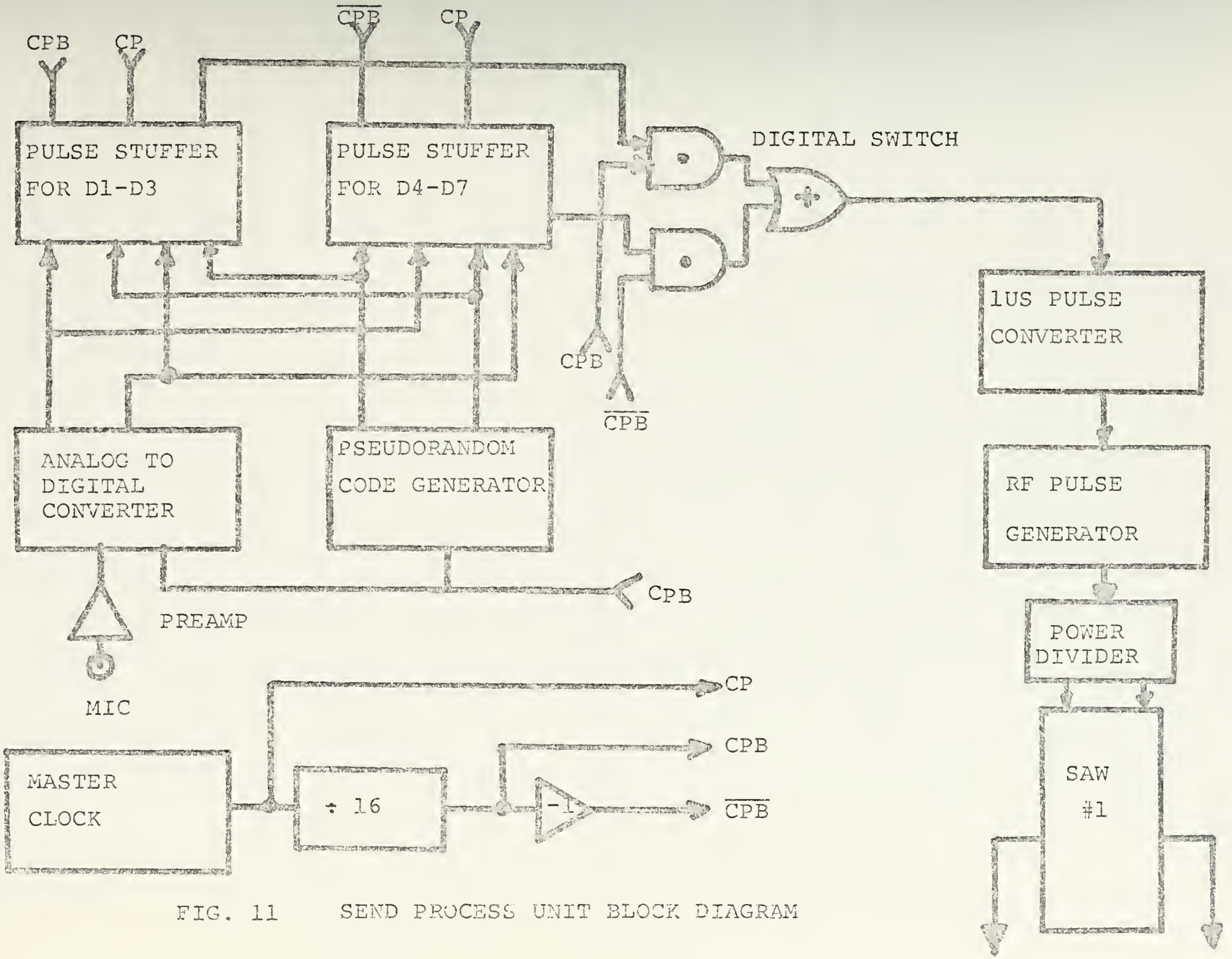


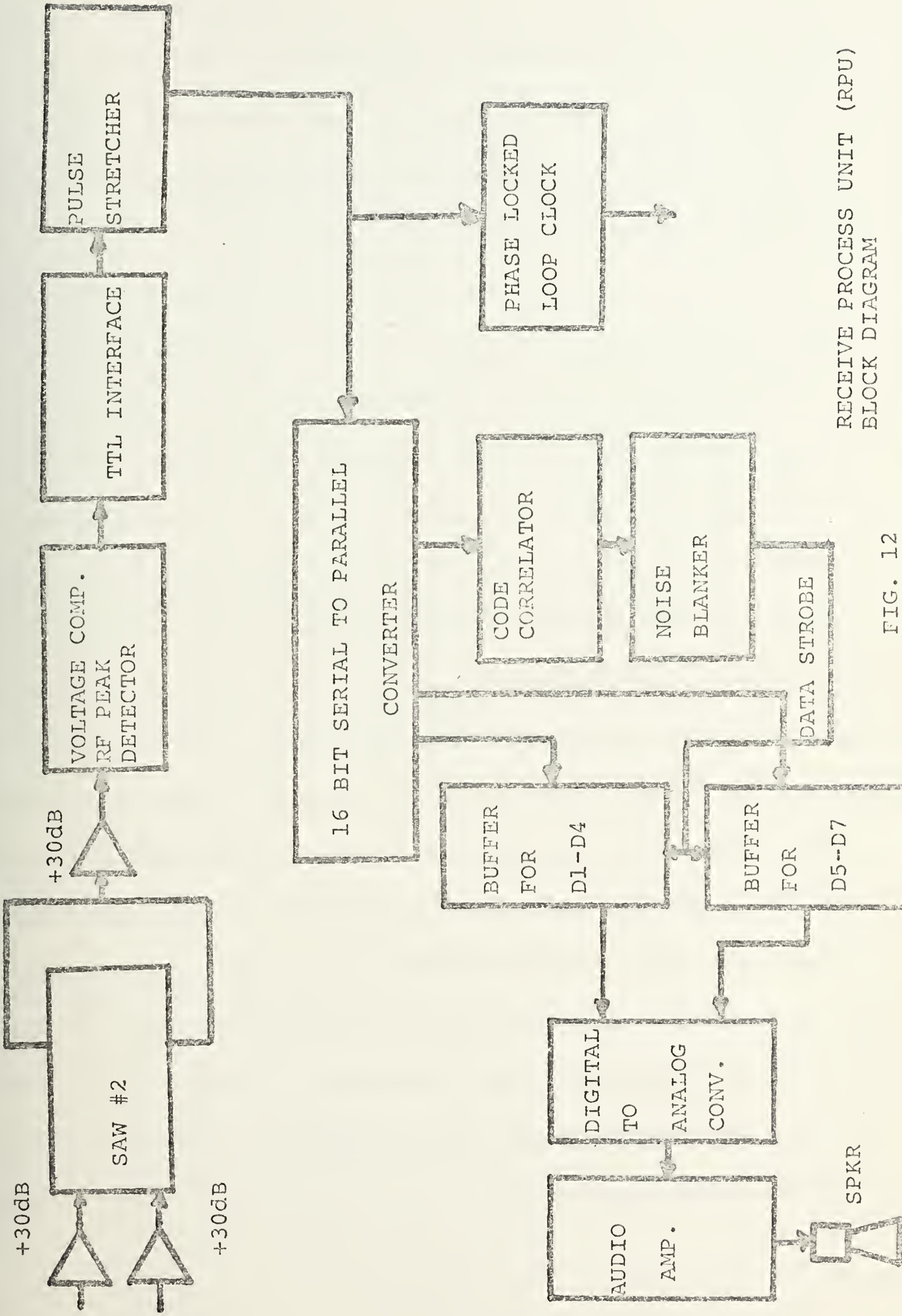
FIG. 11 SEND PROCESS UNIT BLOCK DIAGRAM

The output of the A/D converter goes to the parallel to serial converter which is arranged in two parts of eight bits each. The serial line of each converter bit is connected to the output line by a digital switch. The arrangement is that one eight-bit register is loaded while the other is being shifted.

From the digital switch, the output is a series of square pulses varying in length from 4 to 12 microseconds when the clock rate is 250KHz. These pulse widths are too long to be fed directly to the SAW device as its optimal pulse width is one microsecond. A one microsecond pulse generator was thus added to give a single one microsecond pulse for every master clock pulse as long as the output of the digital switch was high. The SAW transducer was thus furnished pulses of optimum duration.

The one microsecond pulses go to one input of an analog switch. The other input was RF from a crystal oscillator. The DC pulses turn the switch on for the duration of the pulse thus allowing RF to pass for that time. The RF generator is set to operate with a carrier frequency near the center of the SAW passband. Since the SAW device was built to operate at 21.4MHz and had a bandwidth of 1MHz, the RF pulse carrier frequency had to be within $\pm$ .5MHz of the center frequency.

Figure 12 depicts a block diagram of the receive process unit (RPU). The transmitted signal was coupled to the receiver by two channels consisting of coaxial cables. The input signal was amplified



RECEIVE PROCESS UNIT (RPU)
BLOCK DIAGRAM

FIG. 12

by 30dB per channel to help make up for the transmitter SAW insertion loss of 43 dB. From the amplifier, the signal was applied to the second identical SAW device which acted as a matched filter. The outputs of the matched filter were algebraically added with a wired OR and given an additional 30dB of amplification to counter the SAW insertion loss. The signal was next input to an RF peak detector and a pulse stretcher which converted each input pulse to a pulse whose width was equal to the minimum pulse length from the digital switch. The pulse stretcher was a 99% duty cycle retregerable flip-flop. The off period between consecutive pulses is so short that it is not noticed in either the next stage, which is a serial to parallel converter, or the sync input to the phase locked loop clock. The converter output is split with the pins corresponding to the information data connected to buffer memory and the pins corresponding to the P/N umbrella code connected to a correlator. When all bits are in place for one data frame, correlation occurs and an output pulse is seen from the correlator. This pulse is used as a data strobe for the buffer memory. When the strobe reaches the buffers, the data present on the input pins of the memory is entered. Since false correlations do occur within each frame dur to bit placement and noise, a noise blanker was necessary to prevent these false correlations from entering noise into the buffer. The buffer storage output is input to a digital to analog converter and the output of the converter to an audio amplifier.

B. DETAILED SYSTEM OPERATION

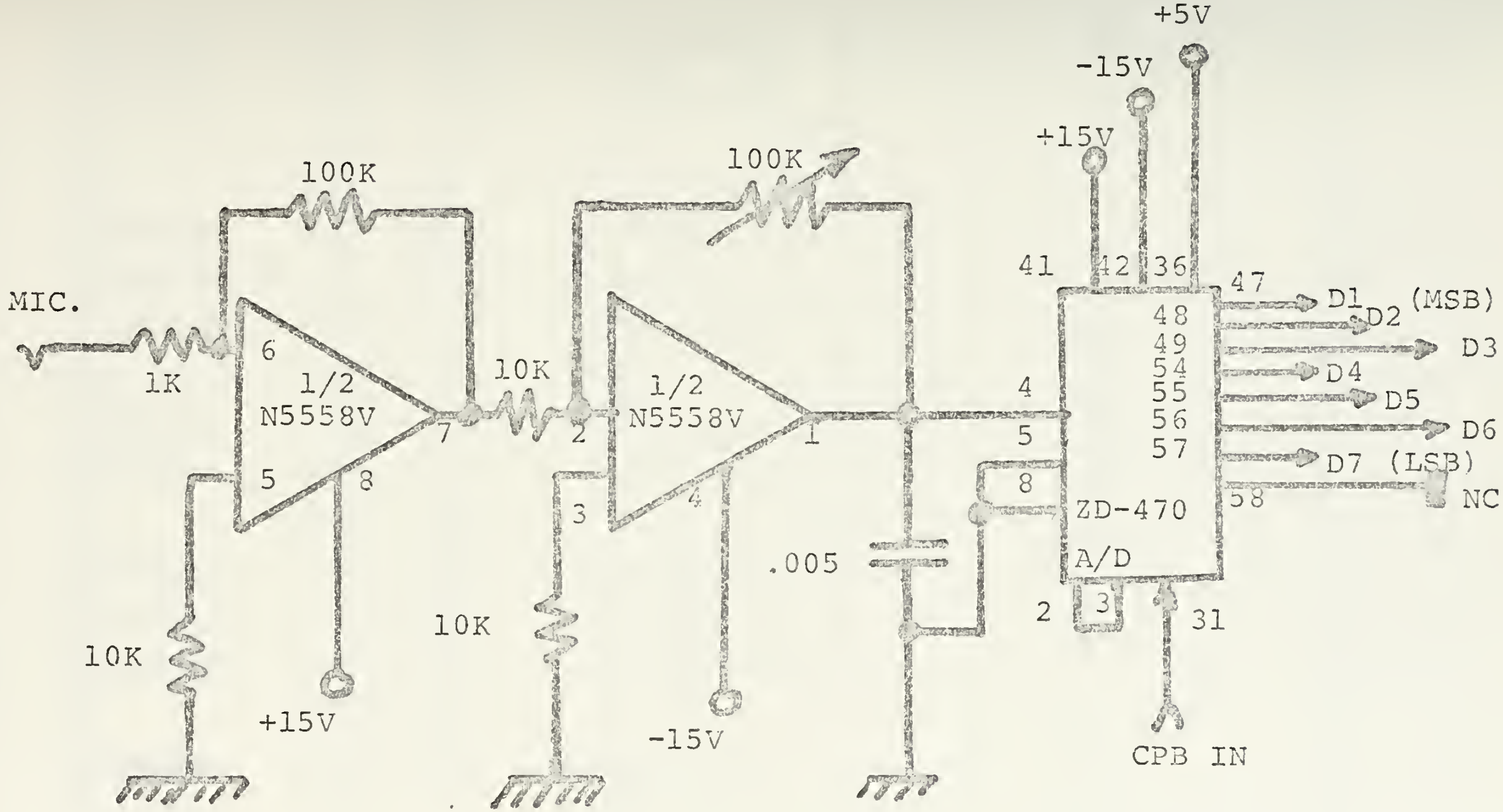
This section presents a component by component description of circuit operation. Waveshapes and spectra are included when helpful for understanding. Good digital design calls for V_{cc} to be bypassed at the chip entry point. As this was done in every case, those capacitors will not be seen in the schematics.

1. Send Process Unit

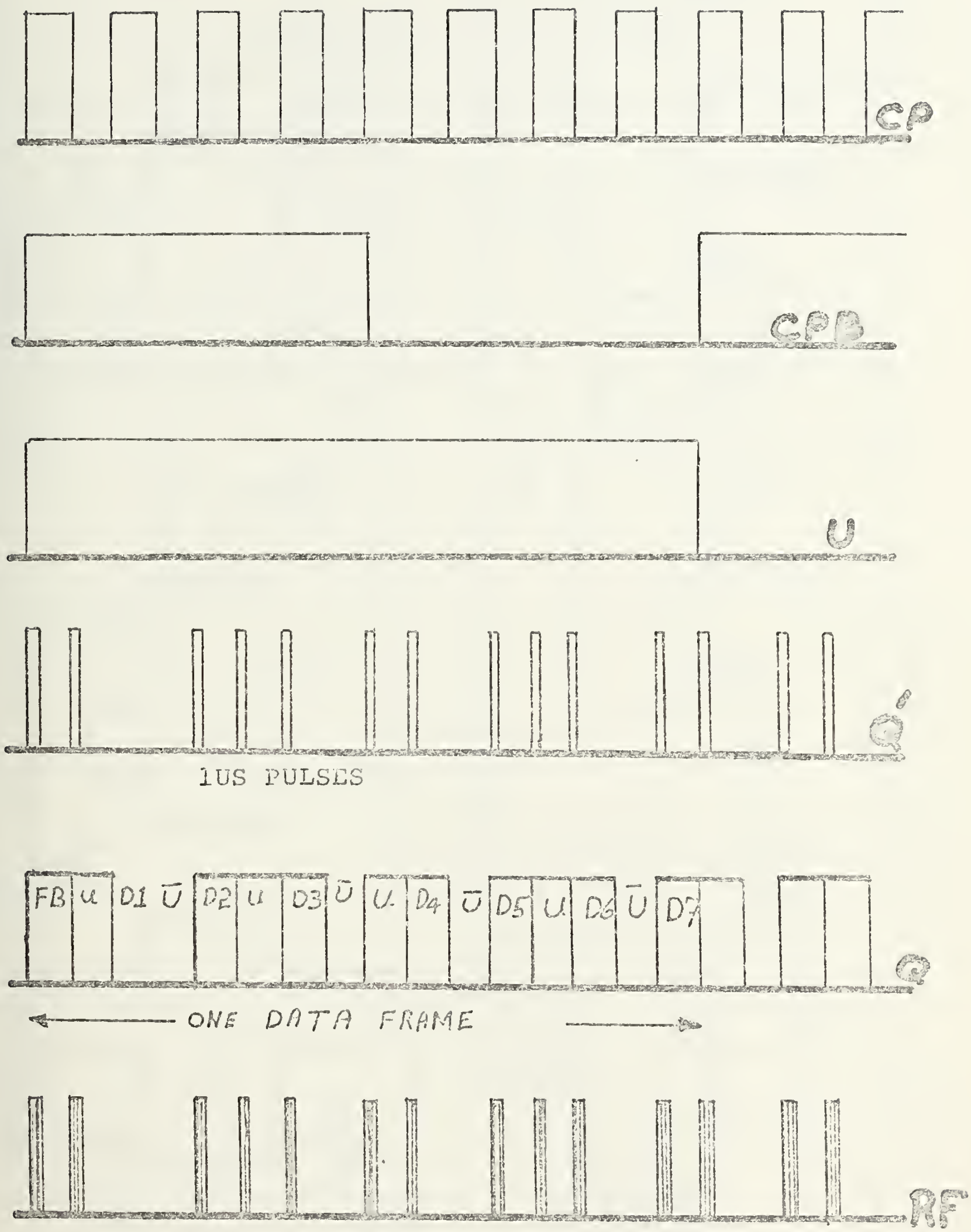
The first block of components (Fig. 13) is the analog preamplifier. The A/D converter required from .25V to 2.0V peak input voltage for proper operation which levied the requirement for a preamp from the microphone input. The entire preamp is an N5558V dual operational amplifier packaged in an eight pin chip. The 1K input resistor and the 100K feedback resistor give the first stage a gain of 100. The DC coupled second stage is variable in gain to a maximum of 10. Fixed high frequency rolloff was provided by a .005UF capacitor across the preamp output.

The A/D converter samples the preamp output at a rate 1/16 that of the master clock rate. Each sample is converted to an eight-bit code word of which the first seven bits are used.

Before continuing with the signal flow, a brief discussion of the SPU timing sequence would be constructive. (Fig. 14) The master clock is designed to operate from 125KHz to 300KHz with a 50% duty cycle. (Fig. 15) A two-phase clock operating at 1/16 of the master



SPU AUDIO PREAMP AND ANALOG TO DIGITAL CONVERTER FIG. 13



SEND PROCESS UNIT (SPU) TIMING DIAGRAM

FIG. 14

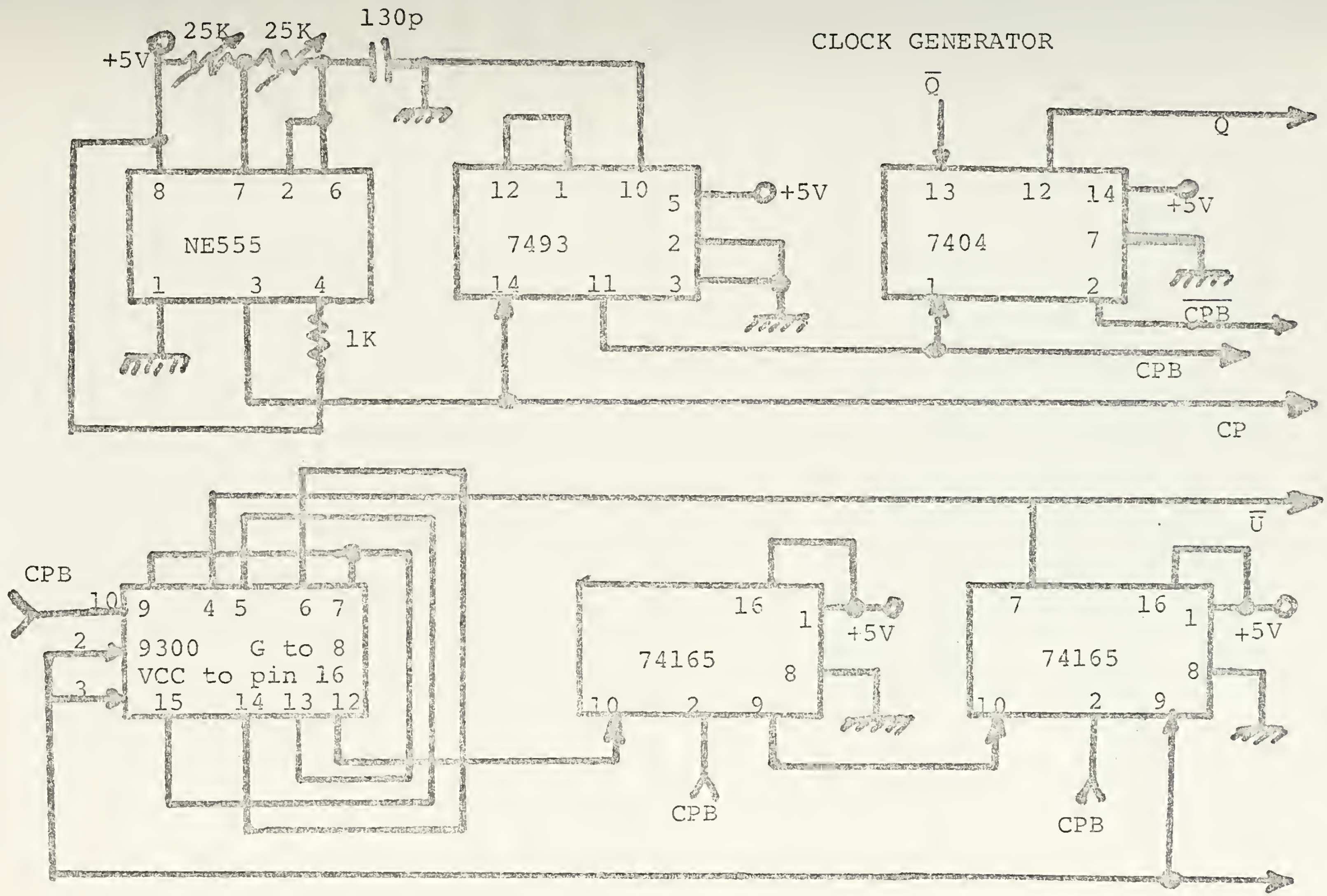


FIG. 15 P/N SEQUENCE GEMERATOR

clock was also necessary for digital switch and load sequencing. In order that the P/N code period be maximized, the P/N generator was clocked at the 1/16 rate. Figure 16 shows the typical P/N output. The master clock frequency was divided by 16 and inverted to give two phases out. The output of the divide by 16 is called "CPB" and is shown in Fig. 17.

The output of the data switch is called "Q", and is longer in duration than the optimal pulse for the SAW devices. "Q" was therefore converted to a series of one microsecond pulses which are optimal for the SAW devices. The pulses are produced at the rate of one pulse for every master clock pulse as long as Q is high.

The P/N sequence generator consists of a 20 bit shift register composed of one 9300 with maximal length code feedback and two eight-bit 74165 shift registers. The generator provides a two-phase output without the need for an external inverter. The pulse lengths of 100Usec shown in Fig. 16 are due to the master clock frequency of 160KHz.

The inputs to the parallel to serial converter are staggered between a data bit from the A/D converter and either U or $\overline{U}$, except for the first bit which is tied high to act as a frame bit. The two parallel to serial converters work as a unit with the digital switch. Clocking of the units is arranged so that one converter is being loaded while the other is being shifted. While CPB is low, the leftmost converter (74165) is being loaded and the clock input (CP) is inhibited. (Fig. 18) The output of the

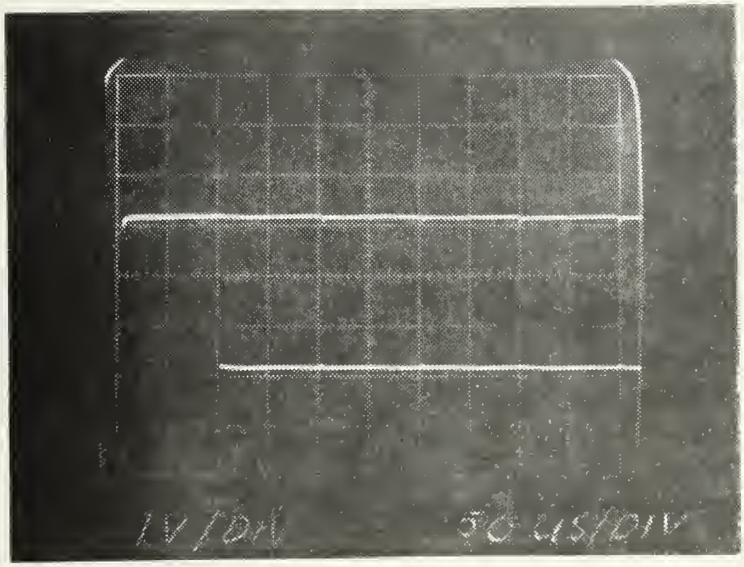


FIG. 16

U P/N OUTPUT

CPB

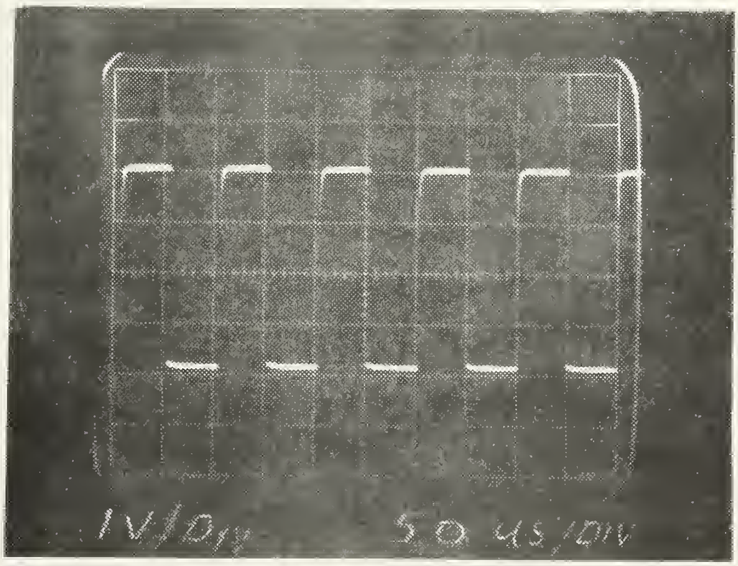


FIG. 17

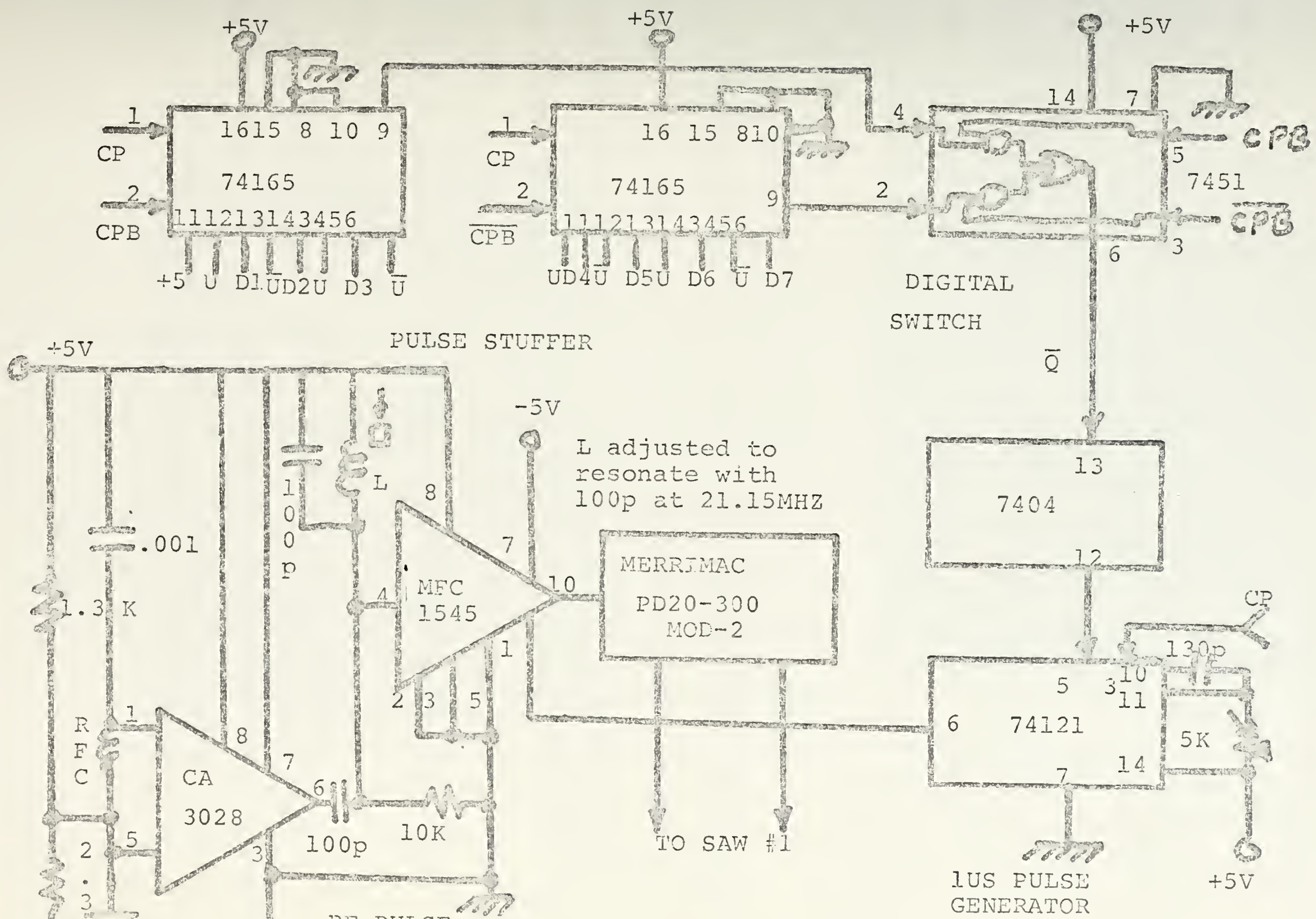


FIG. 18
SEND PROCESS UNIT (SPU)

AND gate in the digital switch is also inhibited by CPB. When CPB is low, $\overline{CPB}$ is high and the previously loaded data in the right hand converter is clocked out with eight pulses of CP before $\overline{CPB}$ goes low again thus inhibiting shifting and initiating another load cycle. When $\overline{CPB}$ is high, the AND gate for the right hand 74165 is enabled allowing the clocked output of that chip to pass through the switch. Sixteen pulses of the master clock will, therefore, clock out one frame of data; eight bits from the left converter and eight from the right. The frames are stacked back-to-back in sequence with this method. The digital switch output, Q, is shown in Fig. 19.

The RF pulse generator board contains the oscillator, one microsecond pulse generator, and the RF pulse generator. The first in the signal flow path is the one microsecond pulse generator which is a monostable non retriggerable flip flop with a built in AND gate at the input that will allow triggering only when both gate inputs are high. The Q pulses are led to one input and the master clock to the other. Whenever Q is high, the clock will trigger the flip flop to produce a 1Usec pulse for each clock pulse.

The RF is generated in a simple op-amp crystal controlled Miller oscillator. DC bias to the input pins is provided by the 1.3K and 2.3K divider. The RF choke is 15 turns of #36 wire wound on a .25 inch toroid.

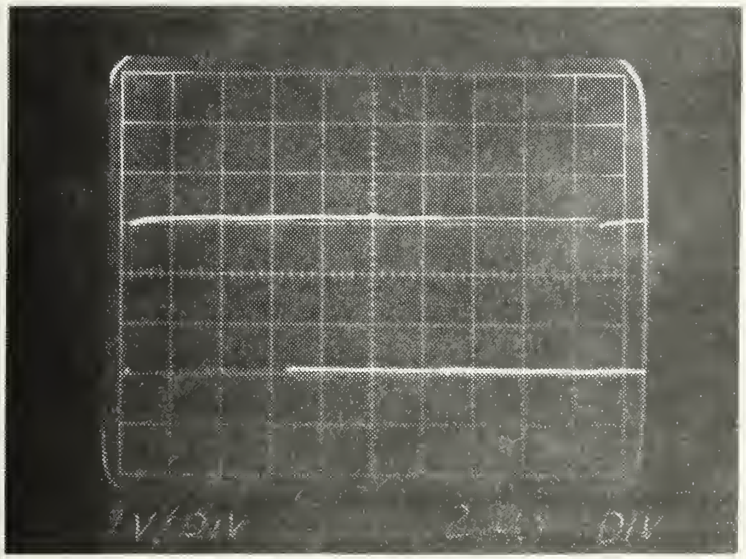


FIG. 19

Q

OUTPUT OF RF PULSE GENERATOR

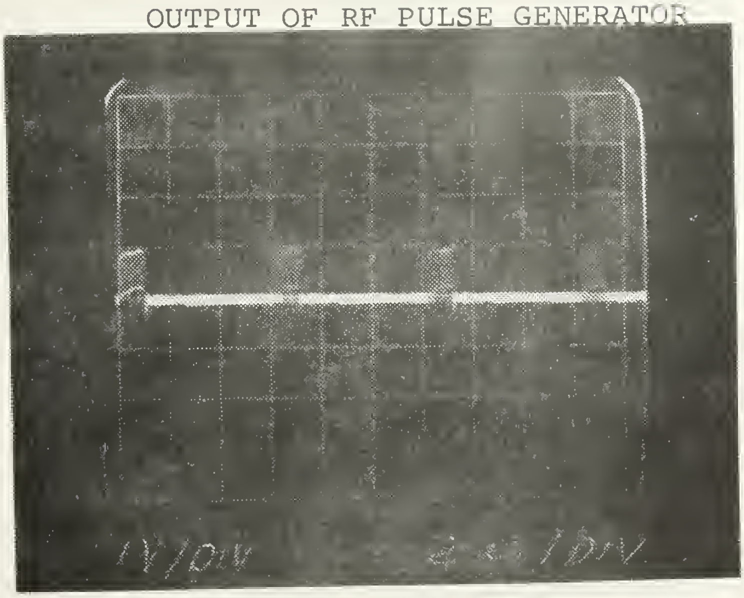


FIG. 20

The analog switch is an MFC 1545 multiplier. One input pin is connected to the oscillator output while the other is connected to the one microsecond pulse generator. Since RF is always present as an input, the output is DC pulses multiplied by steady RF which results in pulsed RF. Figure 20 illustrates the RF pulse output from the switch. This output is split by a power divider for use in the SAW devices.

The output of the analog switch is itself a spread spectrum signal and may be used as such without the Golay sequences (and without the factor of two gain that they provide). Figures 21 through 24 depict various aspects of its spectrum. A good idea of side bands may be seen in Fig. 21. Minor peaks above 40MHz are strong local interference feed through. As may be seen in Figs. 22 and 23, the total spectrum width is 14MHz at the 40 dB down points. The spectrum is not a perfect $(\mathrm{SIN}\ X / X)^2$ but is the spectrum of the trapezoidal RF input pulse. The width of the first or main lobe is twice the inverse of the pulse period of about 2MHz. This is the width predicted by theory. Figure 24 shows the finer structure with minor peaks occurring at 10KHz and 54 KHz intervals. The 10KHz peaks are the result of line spacings at 1/Mt where M is the code period and t is the pulse width. The code period at 160KHz clock rate is 104.86 seconds and t is 1Usec. The result is a spacing of 9.54KHz for the lines. The 54KHz peaks are caused by a pseudo period of 18.75Usec which is the maximum duration of Q and is produced by three one-microsecond pulses occurring sequentially at

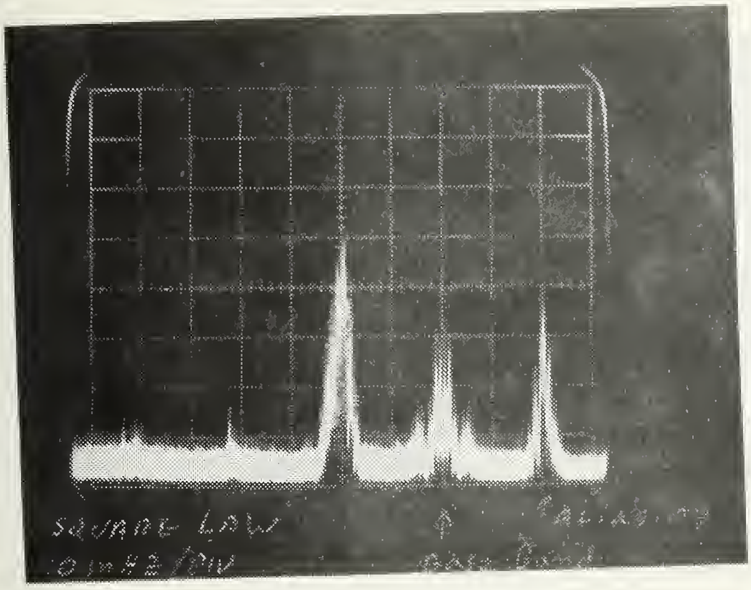


FIG 21

RF PULSE GENERATOR SPECTRUM
SQUARE LAW, 10MHZ/DIVISION

TIME AVERAGE OF RF PULSE GEN
SPECTRUM 2MHZ/DIV, LOG

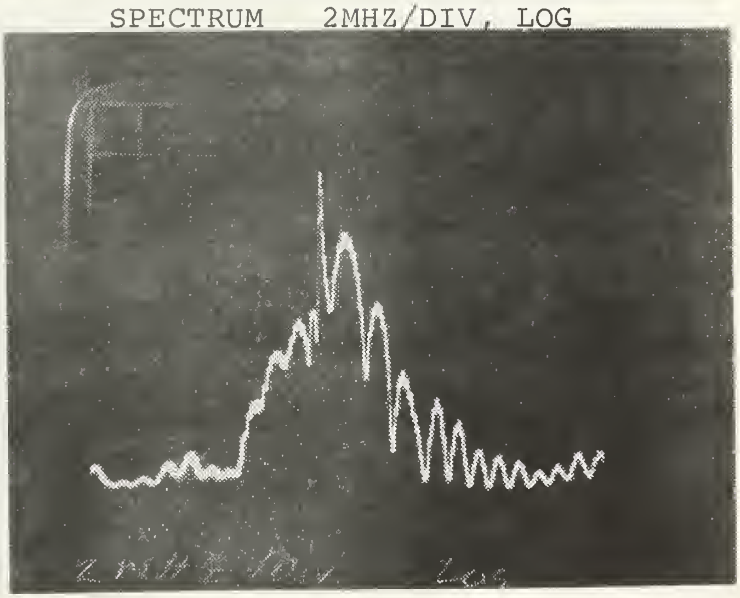


FIG 22

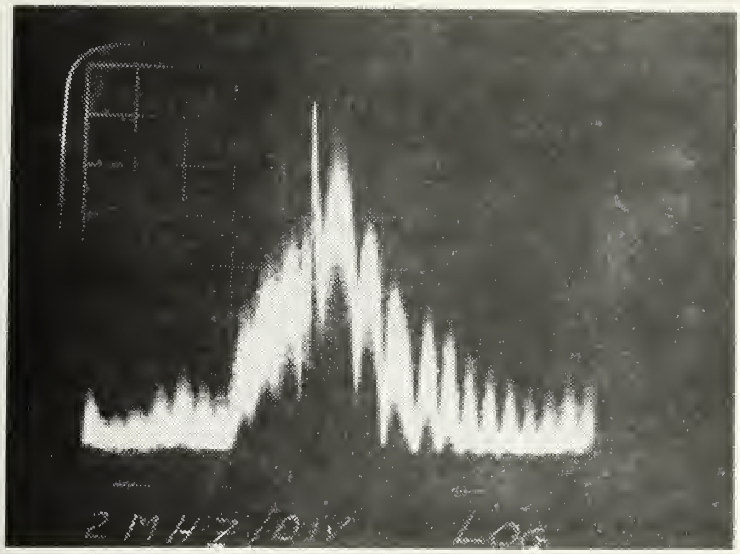


FIG. 23

RF PULSE GENERATOR SPECTRUM
2MHZ/DIV, LOG

SAME AS ABOVE BUT 200KHZ / DIV.

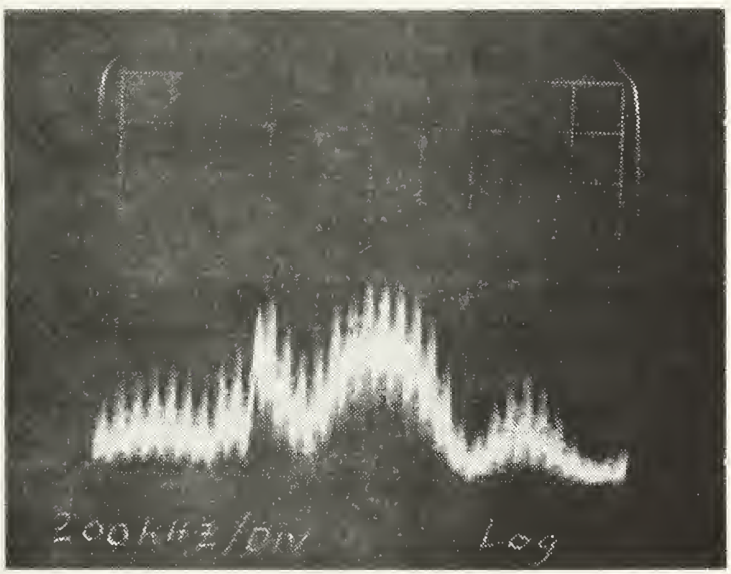


FIG. 24

least once in every frame for the particular 100Hz modulation used for the test.

Figure 25 shows the complex nature of the first SAW device output. Neat triangular waves are not produced because of the trapezoidal input pulse and because the code bit rate is high enough to cause code overlap in the SAW device. Code overlap occurs when one pulse has not reached the end of the finger pair groups before the next pulse arrives and begins to add voltage to the output. With the type of coding used, code overlaps with pulse-to-pulse spacings as close as two finger pair groups can be tolerated. Figures 26 and 27 show the spectrum of one channel of the first SAW device output. Total half power bandwidth is one megahertz. Minor peaks may be seen at multiples of the clock frequency. This undesirable characteristic can be rectified by shaping the input pulses with a $(\text{COSINE})^2$ window. [3]

The outputs of the SAW device in an actual transmitter would now go to frequency conversion and class "A" RF amplification. Class "B" or "C" amplifiers cannot be used as they introduce unacceptable levels of phase distortion, destroying the phase relationships necessary for signal recovery. Any frequency conversion should be accomplished in a double-balanced modulator for the same reason.

Two transmit channels are necessary. If the transmit path is long or if multiple reflections are likely to occur, the frequencies should be chosen as close together as possible so that the phase shift introduced will be approximately the same in each channel.

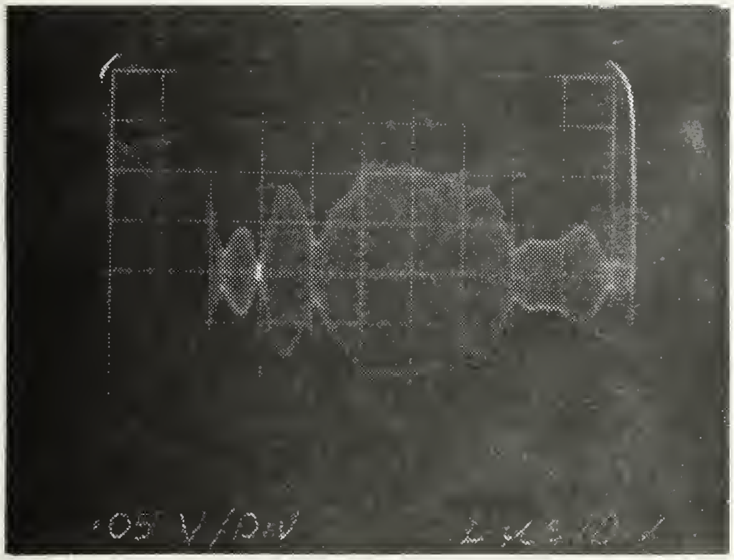


FIG. 25

SAW #1 OUTPUT (+30dB)

SPECTRUM OF ABOVE OUTPUT

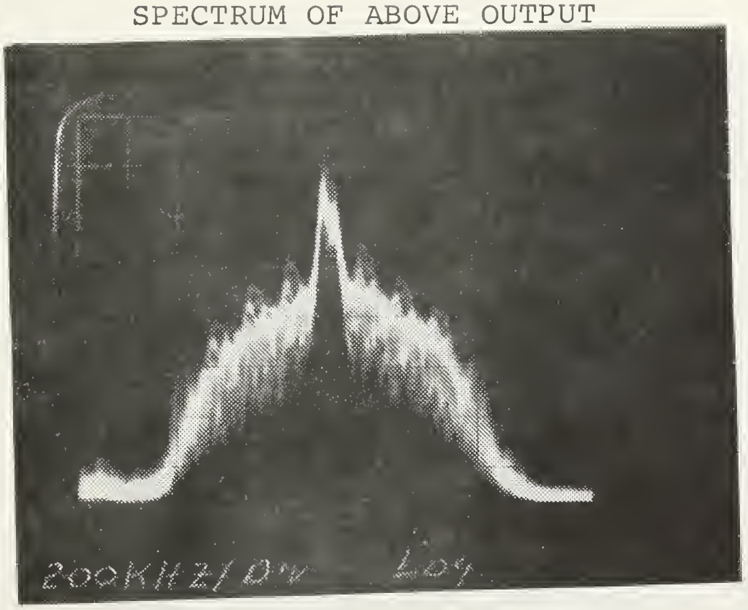


FIG 26

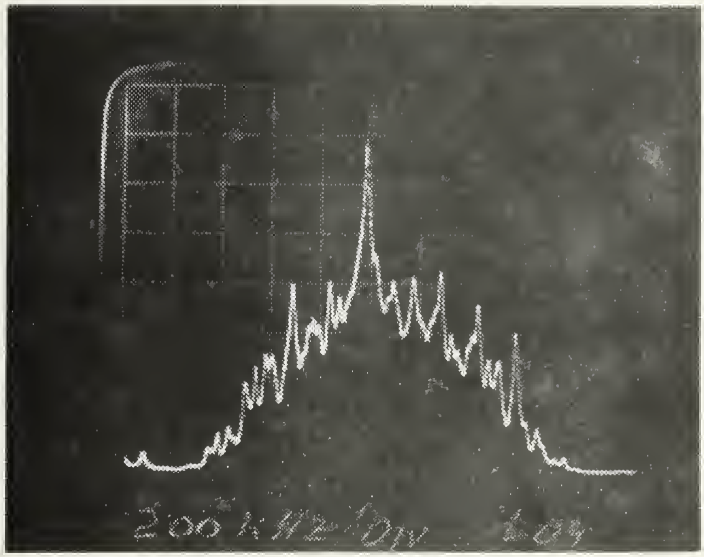


FIG. 27

TIME AVERAGE OF FIG. 26

SAW #1 SPECTRUM AT 500KHZ/DIV.

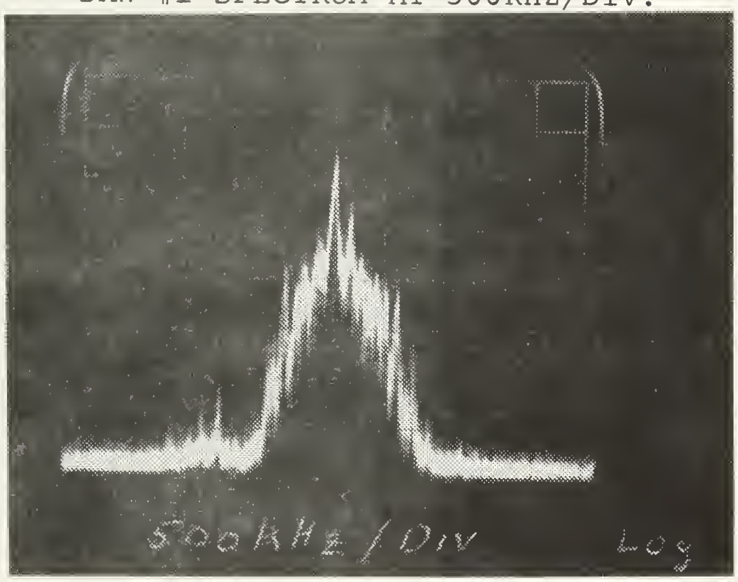
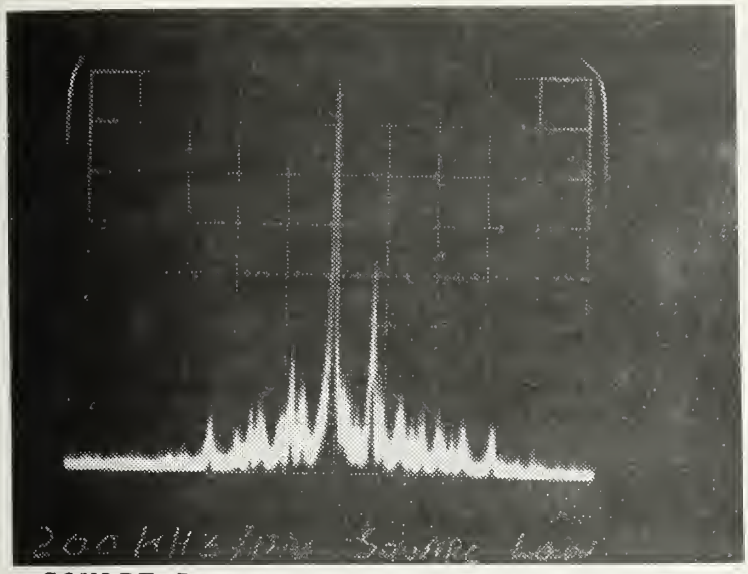


FIG. 28

2. Receive Process Unit

The first SAW outputs are coupled by coaxial cable to the two channel amplifiers of the RPU. These amplifiers are necessary to make up for the SAW device insertion losses. The outputs of the channel amplifiers are sent to the second SAW device operating as a matched filter. The outputs are taken from its phase-coded interdigital transducers and added together in an RF "tee" junction. The output from that "tee" junction is shown in Fig. 30. Fig. 31 shows the spectrum at that point. Amplification is again necessary to make up the insertion loss of the second SAW device before the peak detector can use the signal. The output of the peak detector is shown in Fig. 32.

Before continuing with the RPU operation, a brief discussion of the RPU timing may prove helpful. (Fig. 33) The peak detector pulses must be widened to reconstruct the Q pulse as nearly as possible for use by the serial to parallel converter. The reconstruction device is a retriggerable flip flop with a 4Usec pulse width and a 99% duty cycle at 250KHz bit rate. Four microseconds was chosen to allow optimum operation from 125KHz to 250KHz and degraded operation to 300KHz clock frequency. The operation above 250KHz is degraded since the output pulses of the pulse stretcher are exactly as wide as the Q pulses except the last one in a series which is still 4usec long. The extra length of that last pulse causes it to overlap into the next data bit space. That space will be called a zero by the serial to parallel converter up to 80% overlap. Above that, a false pulse will be entered.

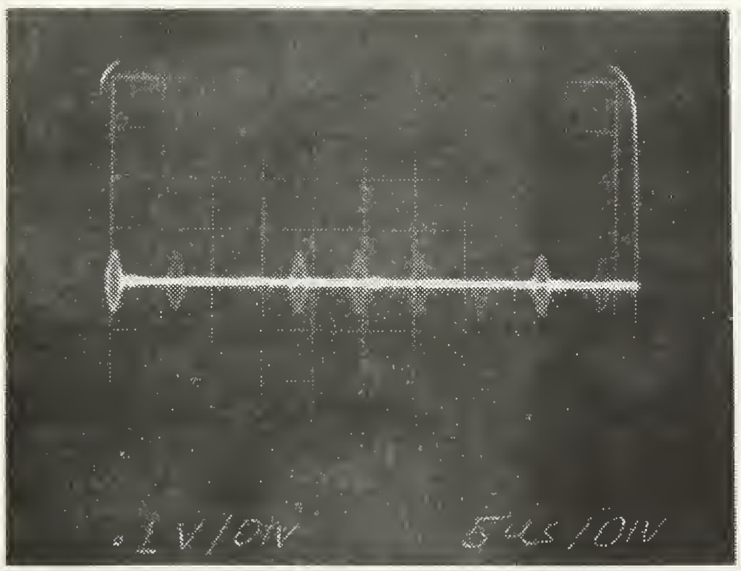


SQUARE LAW SPECTRUM OF SAW #1

FIG. 29

AUTOCORRELATION SUM AT SAW #2

FIG. 30



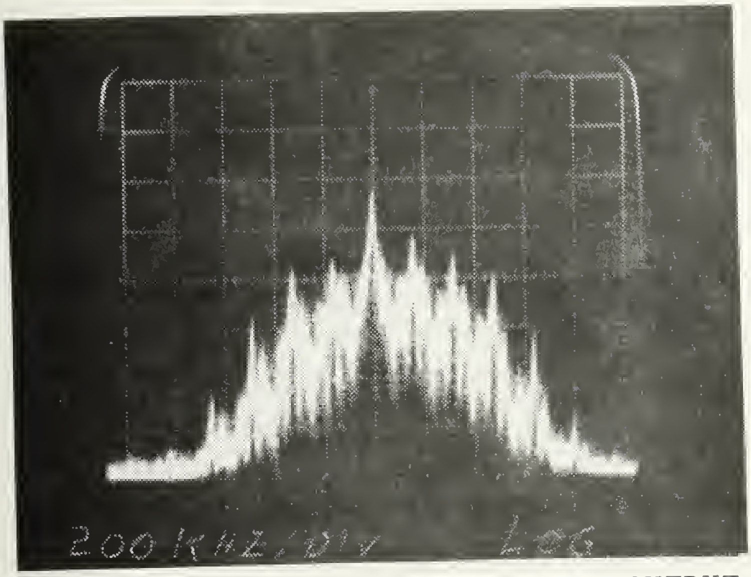
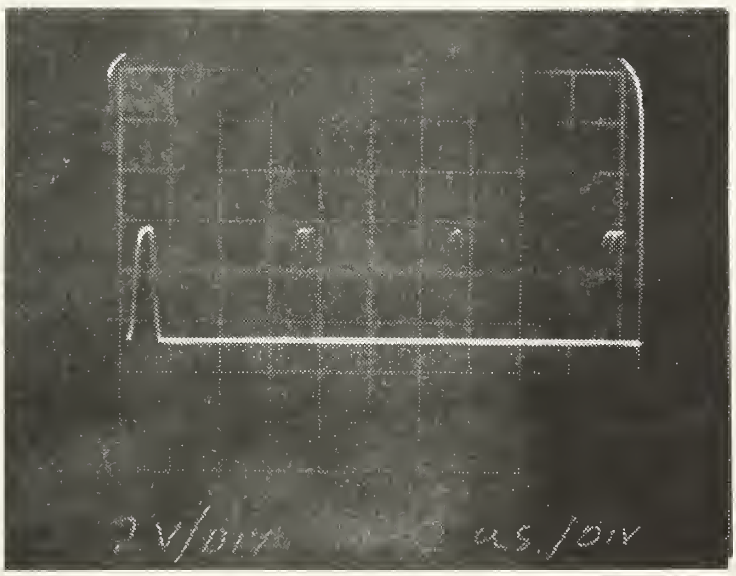


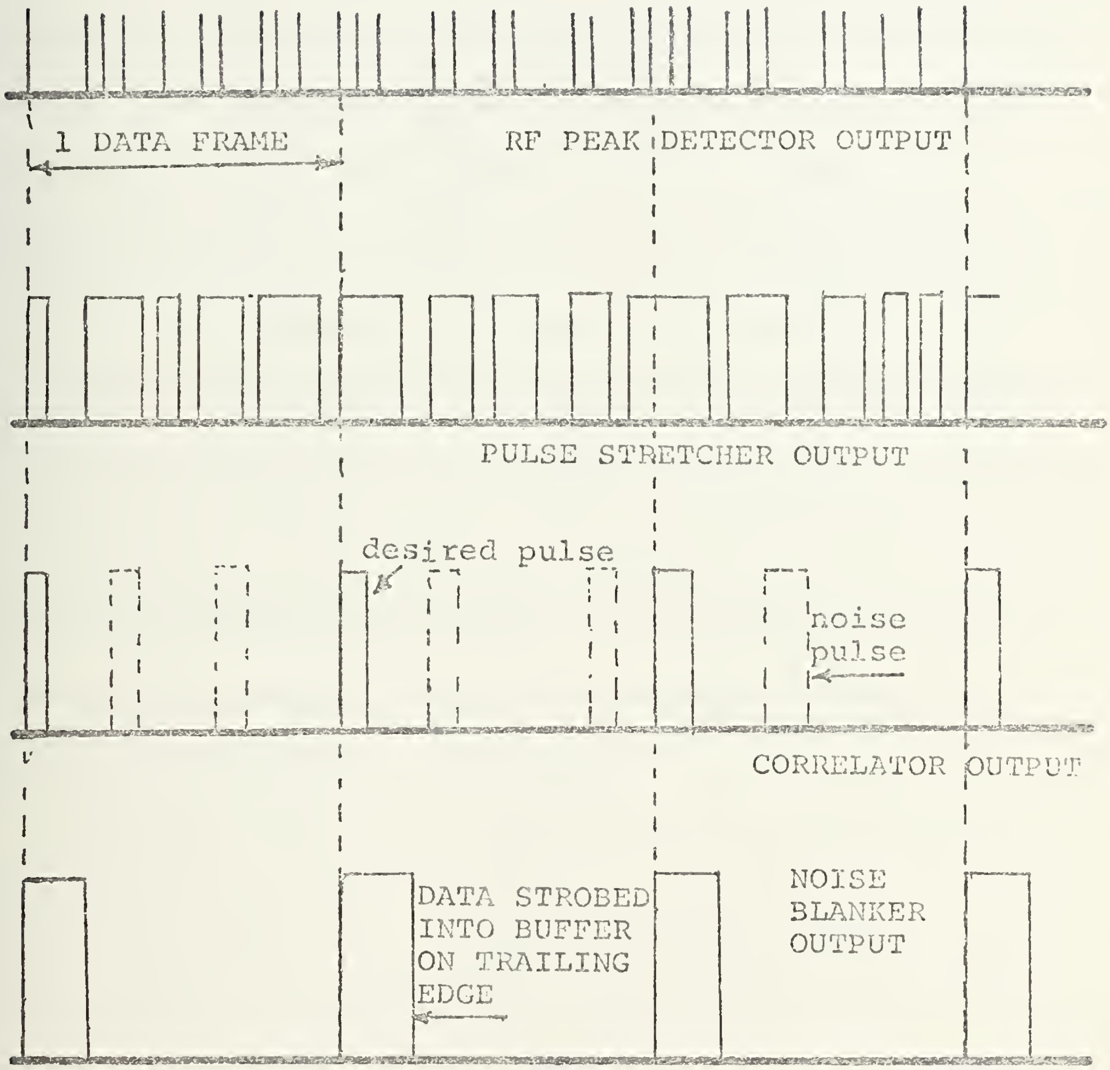
FIG. 31

SPECTRUM OF AUTOCORRELATION SUM OUTPUT

RF PEAK DETECTOR OUTPUT

FIG. 32





RECEIVE PROCESS UNIT (RPU)
TIMING DIAGRAM

FIG. 33

The serial to parallel converter output is wired to four exclusive OR gates for determination of correlation. When the first bit in the converter and the outputs of the exclusive OR are simultaneously high, correlation has occurred and the AND gate to which all are connected puts out a pulse. Unfortunately occasional false correlations occur within each data frame. Since the correlation pulses are eventually to be used to strobe data into buffer memory, to leave the false correlations in would cause a noisy output as meaningless data would also be input to the memory. A noise blanker consisting of a non retriggerable flip flop was added and the output of the noise blanker used as a data strobe. Since the flip flop is not retriggerable, the false correlations do not cause noise to be strobed into memory. A gate delay problem also existed with the correlator, therefore the pulse width of the noise blanker was made adjustable to compensate.

Figure 34 shows the RF peak detector and pulse stretcher circuitry. The leak detector is an analog voltage comparator with a variable threshold set by the 25K pot attached to pin 3. This pot is adjusted to give the peaks previously seen in Fig. 32. The signal input amplitude is high enough to allow a wide latitude in the setting of this control. The input to the voltage comparator from the second SAW device is matched with a series 75 ohm resistor. The reference pin is DC matched to ground through a 68 ohm resistor and is AC bypassed to ground. The output (pin 7) was only .4V peak which was too low for

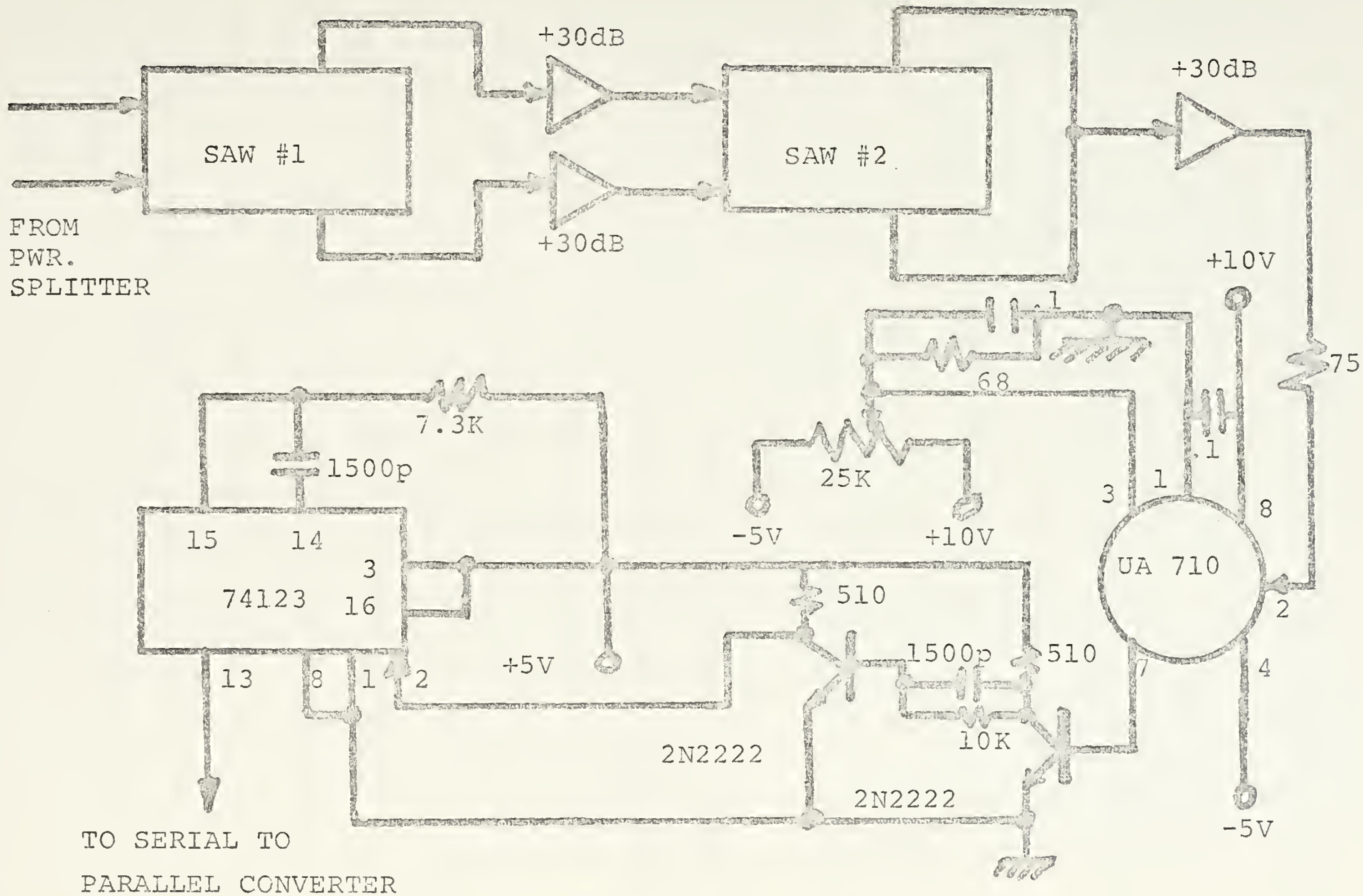


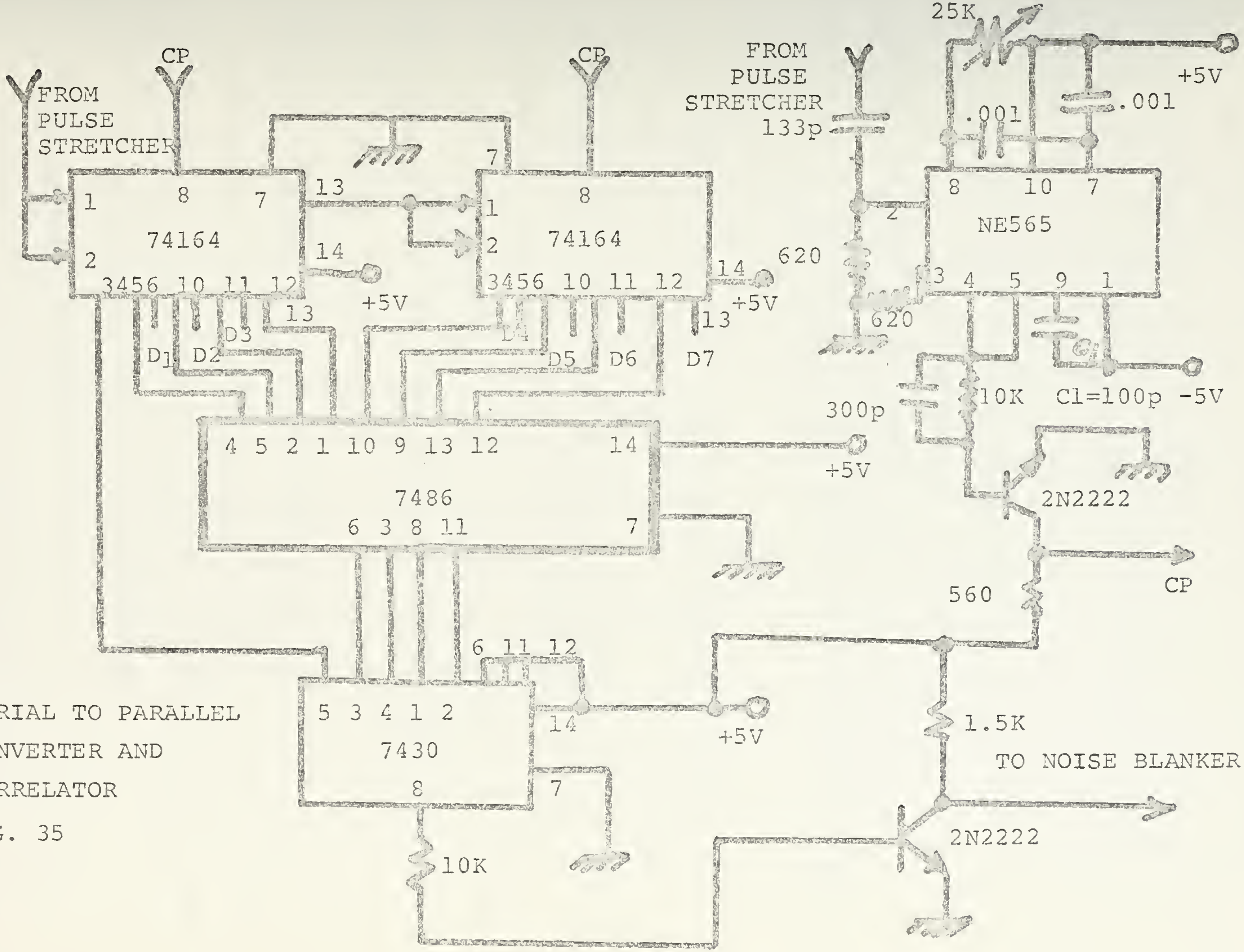
FIG. 34 RF PEAK DETECTOR AND PULSE STRETCHER

TTL compatability. A two transistor pulse amplifier with frequency compensation was added to bring the .4V up to 4.7V. The second stage output was coupled directly to the pulse stretcher flip flop and so will not be discussed further.

The serial to parallel converter input was directly connected to the pulse stretcher output. The converter consists of two eight bit converters in series since at the time of this writing, a single 16 bit converter is not made. The input is clocked through the converter by clock pulses from a phase locked loop. The outputs from the converter (Fig. 35) are connected either to the exclusive OR gates or to the buffer inputs except for the first bit which is a frame bit.

The phase locked loop is an NE565 connected in a conventional manner. In order to reduce the effects of noise pulses, the sync input was connected to the pulse stretcher by a capacitor and resistor differentiator which gives too narrow a pulse width on noise spikes for sync. The output of the NE565 was just within limits for TTL circuitry so a pulse amplifier was added to its output to insure compatibility.

The correlator consists of a quad exclusive OR gate and an eight input NAND gate with an inverter stage following. The noise blanker is a simple non retriggerable flip flop with an adjustable pulse width. In operation, the simplest way to adjust the blanker pulse is to make the width of the pulse equal to the width of the correlation pulse and then shorten it until the noise heard in the speaker drops abruptly.



SERIAL TO PARALLEL
CONVERTER AND
CORRELATOR

FIG. 35

At that point, the gate delays are compensated for and only good data is entering the buffers. Figure 36 shows the schematic diagram of the correlator and noise blanker as well as the buffer memory and D/A converter.

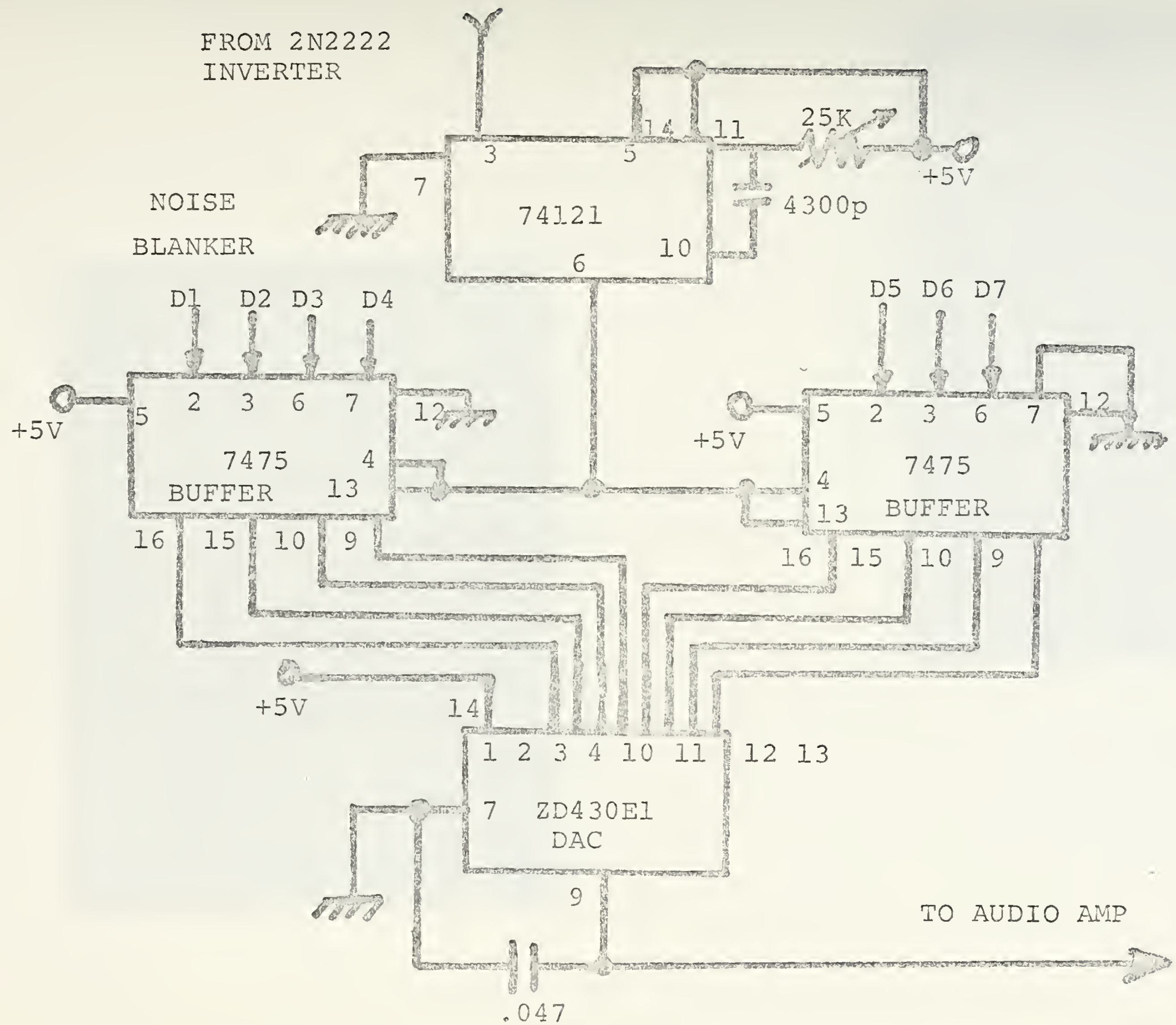
The buffers are two 7475 four bit latches. They accept an input strobe and will input data only when the strobe inputs are high. They latch and store the input data when the strobe pulse goes low. Simultaneously, data is presented at their output pins where it is input to the D/A converter.

After conversion to analog, the signal passes through a frequency compensation network which removes much of the conversion hiss. The signal is then passed to an audio amplifier. Figure 37 shows the noise blanker input and output and Fig. 38 shows the D/A output.

The audio amplifier (Fig. 39) consists of a N5558V preamplifier with a fixed gain of 10 and further high frequency hiss suppression circuits. The final stage is an LM-380 two-watt power amplifier with capacitive coupling to the speaker. Since the LM-380 has a tendency to oscillate under no load at a frequency near 1MHz, a 2.7 ohm resistor and a 1Uf capacitor to ground were added to the output to suppress those oscillations. Figure 42 shows the RPU state diagram.

3. The Completed System

The entire system is uncomplicated and uses standard TTL and analog circuitry in a straightforward manner. The packing density



RPU DIGITAL TO ANALOG CONVERTER AND BUFFERS FIG. 36

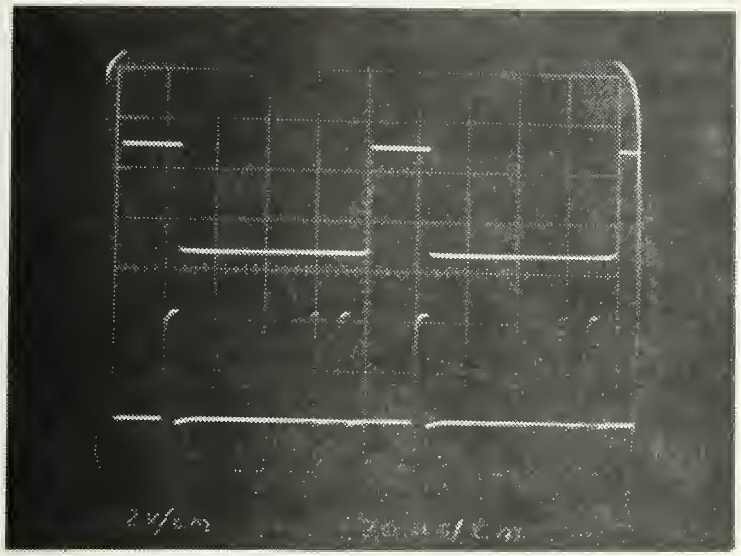
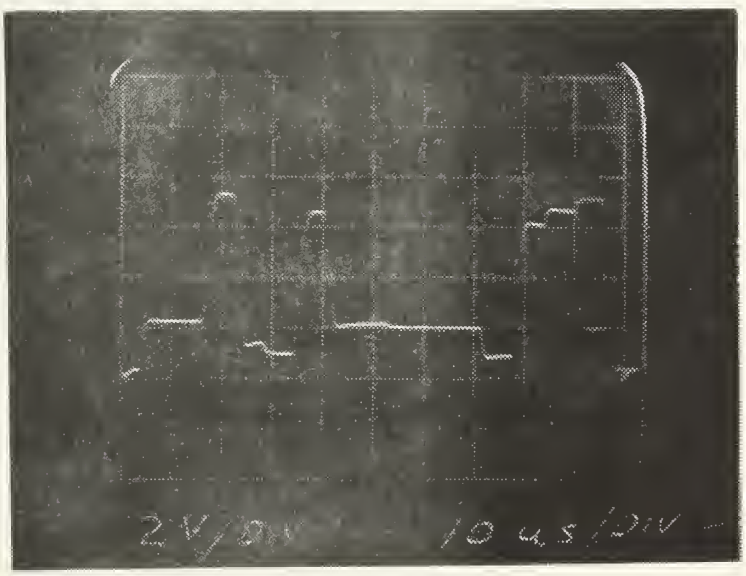


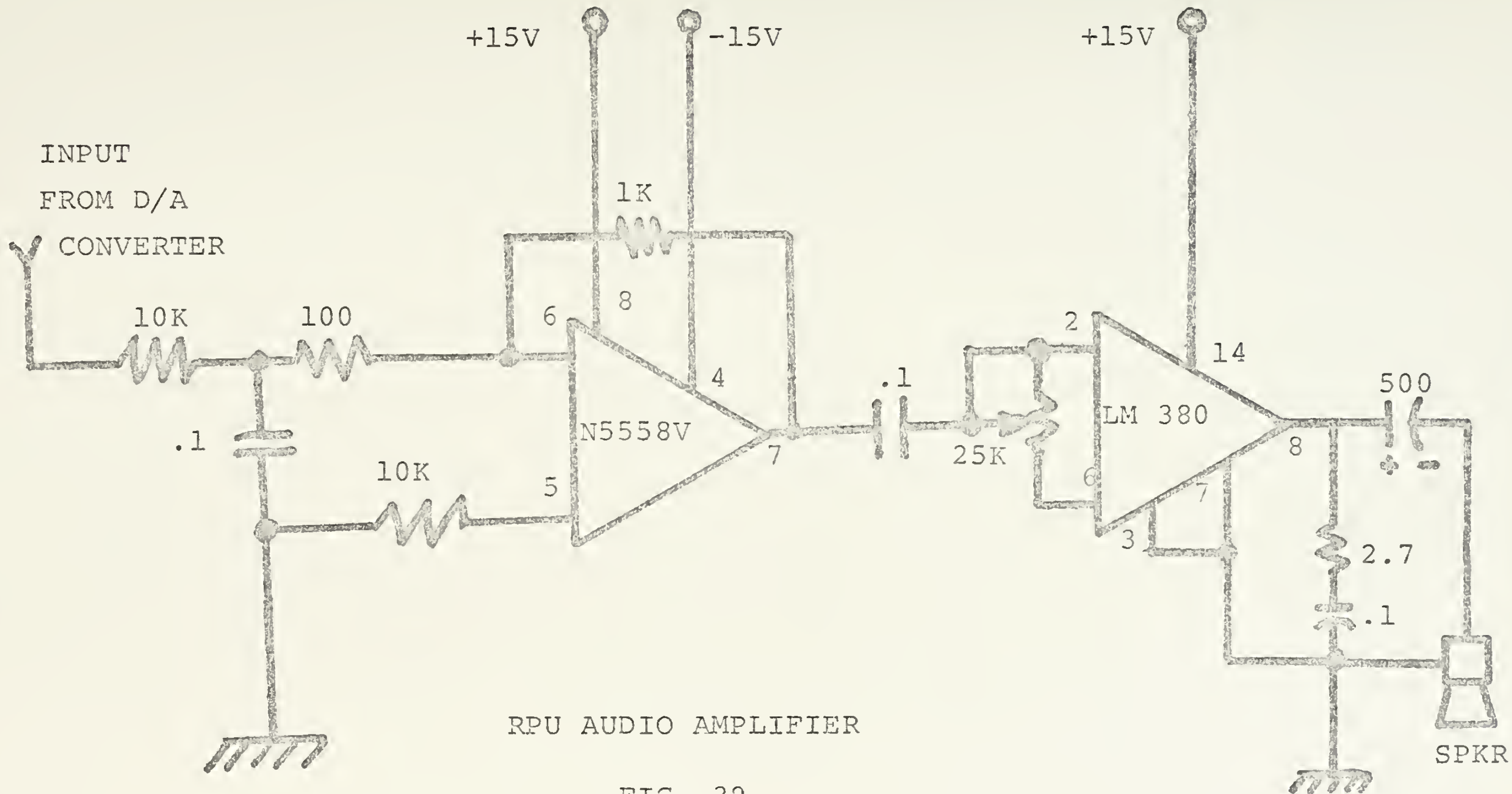
FIG. 37

TOP: NOISE BLANKER OUTPUT
BOTTOM: NOISE BLANKER INPUT

DIGITAL TO ANALOG CONVERTER OUTPUT

FIG. 38





RPU AUDIO AMPLIFIER

FIG. 39

of the laboratory device is low for purposes of testing and ease of construction. The entire package measured 13 by 5 by 3 inches. Higher packing densities as are now used will permit the entire unit to fit inside part of a handheld transceiver case similar to those in use by the fleet for bridge-to-bridge communications.

IV. SYSTEM TESTS

The completed system was tested to determine its resistance to wideband noise and CW interference. Additionally, two methods of loop stealing were tried and the effectiveness measured.

A. NOISE REJECTION TEST

Figure 40 depicts the test setup used for determining the noise rejection capabilities of the system. The noise generator, built gy General Radio, had a spectrum that was flat from 100Hz to 100KHz and that was within $\pm$ 5dB of flat from 4Hz to 5MHz. The noise was up-converted to double sideband suppressed carrier by means of a URM-25 signal generator and a double-balanced mixer. Power measurements were taken with an HP-431 wideband power meter and a single pole double throw coaxial switch to change the output of the double balanced mixer from the circuit to the meter. The noise was inserted into one channel of the RPU channel amplifier input. The effect was qualitatively measured by listening to the speaker for zero intelligibility of a tape recorded message. Quantitative measurements were taken with the aid of a Techtronix 442 oscilloscope connected to the correlator output and the noise blanker output. The point at which the interference caused the noise blanker to produce a spurious output was noted as the point of first effect on the circuit. The effects on loop lock were quantitatively

determined by using a two channel counter, the USM-245A, which was set for a ratio of frequencies. One frequency was the master transmit clock while the other was the loop clock. The loop VCO frequency was moved so that the loop was just on the border of its capture range when in lock. The ratio was 1.0 when the loop was in lock. As the loop lock degraded, the ratio changed to indicate the amount of frequency difference. Since the capture range of the loop was only 40% of the clock frequency, complete break of lock was recorded when the counter read the ratio of .60. In most cases the point of zero intelligibility was very near the point of drop lock. The results of the single channel noise test are listed in Table 1 on the following page. Frequencies for the noise center frequency were chosen around the SAW center design frequency of 21.4 MHz. The signal power was the power measured at the SAW disperser output by a method described at the end of this section. The signal to noise ratio used was the signal to the noise added at the point of zero intelligibility. That value of noise was felt to have more significance than some other values as it indicated a limit to the design. The "noise" here, then, is the interference and not thermal or natural noises.

The effects of the interference were not symmetrical about the center frequency as the spectrum was not symmetrical due to the trapezoidal input pulse. Examination of Table 1 will show that even near the SAW center frequency large powers relative to the signal

Freq. MHz	Signal dBm	First Effect dBm	Zero Intell. dBm	Drop Lock dBm	Signal to Zero Intell. Noise ratio dBm
16	-56.3	-1.4	3.8	9.7	-60.1
18	-56.3	-1.6	0	0.5	-56.3
19	-56.3	-2.5	-2.0	-2.0	-54.3
20	-56.3	-16.4	-11.2	-9.8	-45.1
21.4	-56.3	-21.2	-20.0	-18.1	-36.3
22	-56.3	-15.6	-15.2	-14.3	-41.1
23	-56.3	-0.7	0.7	1.7	-55.6
24	-56.3	1.2	4.3	9.1	-60.6
26	-56.3	5.7	7.1	14.0	-66.3

TABLE 1

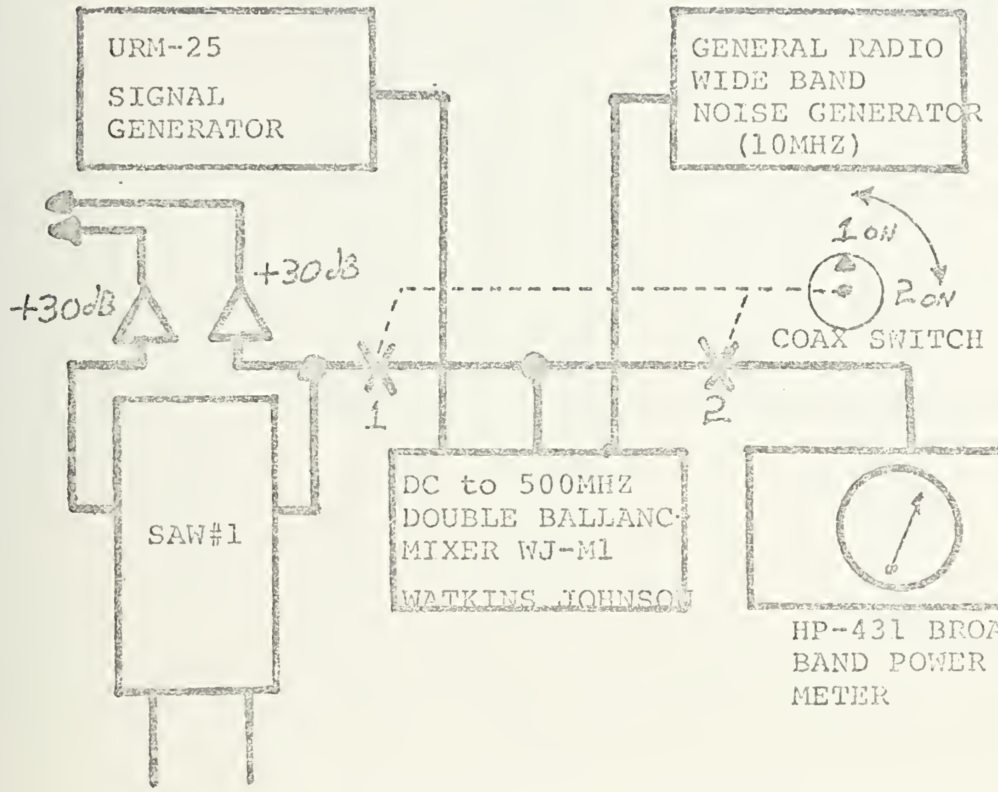
SINGLE CHANNEL NOISE REJECTION TEST

power are necessary to cause an effect in the output. This is one of the advantages of spread spectrum as previously discussed. The effect becomes even more pronounced as the jammer center frequency is removed from the SAW center frequency.

In addition to the single channel test, a two channel noise test was also performed. The up converted noise was added to both channel amplifier inputs. The results of this test are shown in Table 2. Comparison of Tables 1 and 2 will show that approximately double the power is necessary to jam both channels. This is to be expected. The results of the two channel test follow the same pattern as those in Table 1. Since the system will work marginally with only one of the two channels a method of detecting single channel jamming might be devised and the jammed channel turned off. The noise would thus be prevented from entering the system.

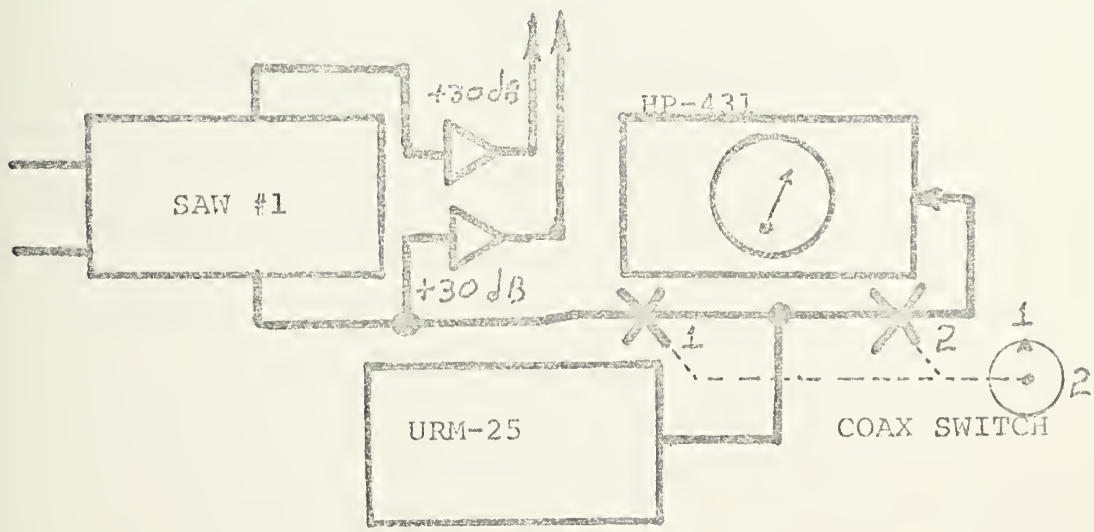
B. THE CW REJECTION TEST

The system was next tested for its ability to reject a CW interference. The test setup is as indicated in Fig. 41. The equipment used for quantitative measurements was the same and in the same configuration as that used in the noise tests. The results of this test are shown in Table 3 on the next page. Because of the similarity to the noise test, only a single channel test was performed. The pattern of the results again followed those of the first two tests.



TEST SETUP FOR NOISE REJECTION

FIG. 40



TEST SETUP FOR CW INTERFERENCE REJECTION

FIG. 41

Freq. Mhz	Signal dBm	First Effect dBm	Zero Intell. dBm	Drop Lock dBm	Signal to Zero Intell. ratio	Noise dBm
16	-56.3	-4.3	0.4	6.7	-56.7	
18	-56.3	-4.5	-2.8	-2.1	-53.5	
19	-56.3	-5.1	-5.1	-5.1	-51.2	
20	-56.3	-19.2	-14.0	-12.5	-42.3	
21.4	-56.3	-23.0	-22.0	-21.2	-34.3	
22	-56.3	-18.2	-18.1	-17.0	-38.2	
23	-56.3	-3.5	-2.2	-0.9	-54.1	
24	-56.3	-1.7	1.2	6.0	-57.5	
26	-56.3	2.6	3.9	9.8	-60.2	

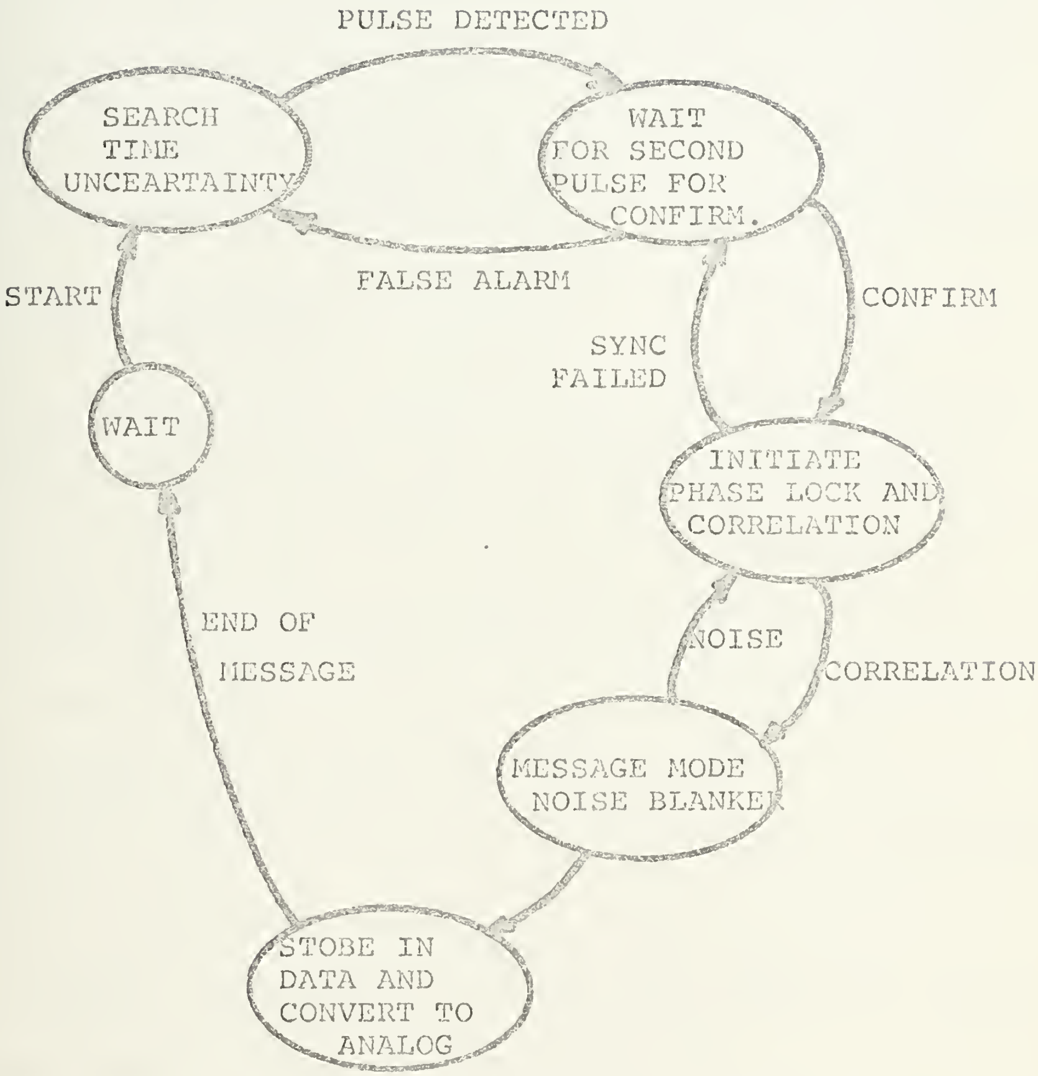
TABLE 2

TWO CHANNEL NOISE REJECTION TEST

Freq. MHz	Signal dBm	First Effect dBm	Zero Intell. dBm	Drop Lock dBm	Signal to Zero Intell. Noise ratio dBm
16	-56.3	-6.0	-4.7	-4.5	-51.6
18	-56.3	-5.3	-2.3	-2.3	-54.0
19	-56.3	-11.2	-4.8	-4.0	-51.5
20	-56.3	-19.5	-16.2	-15.7	-40.1
21.4	-56.3	-29.9	-20.7	-19.9	-35.6
22	-56.3	-20.5	-20.2	-18.2	-36.1
23	-56.3	-3.4	-0.5	-0.5	-55.8
24	-56.3	-0.4	0.1	1.8	-56.4
26	-56.3	3.9	5.8	6.8	-62.1

TABLE 3

SINGLE CHANNEL CW INTERFERENCE REJECTION TEST



STATE DIAGRAM FOR DEMODULATOR

FIG. 42

Comparison of Tables 1 and 3 shows that the system is much more sensitive to CW interference than to noise interference. This is because the wider the input signal in bandwidth, the wider the resultant output from the despreader since it spreads the interference. Thus, the wider the interference, the less effect it has on the system since its spectral power density is lower than that of the de-spread desired signal unless large powers are used to cause interference. [3] The most effective interference, therefore, is CW when applied to a direct sequence system.

C. TWO LOOP CAPTURING TECHNIQUES

Table 4 illustrates what can happen when the system sync depends only on clock sync and not code sync. Two methods of capturing the loop were successfully tried. The first of these was RF swept at the clock rate. The sweep width was gradually increased until further increases brought diminishing returns. The second method was a CW carrier amplitude modulated with a swept audio signal centered on the clock frequency. The carrier was exactly at the SAW design frequency. In both cases, the signal was strong enough to leak through the SAW device correlator. The input to the system in both methods was at the RPU channel amplifier input as in the rejection tests.

In method one, as the RF was swept past the SAW bandpass, a pulse was formed by leaking RF which was detected by the peak detector and treated as if it were an actual signal pulse. The signal pulses were

	METHOD I	METHOD II
Center Frequency (MHz)	21.4	21.4
Power Required to Capture the Loop (dBm)	-20.5	-21.0
Signal Power (dBm)	-56.3	-56.3
Sweep Range	20 - 22.5 MHz	50 - 200 KHz
Loop Pull Off (KHz)	110 - 230	110 - 230
Clock Frequency (KHz)	170	170

TABLE 4

TWO LOOP CAPTURE TECHNIQUES

of lower amplitude at the power level of loop capture and so were not effective. Loop capture was determined by setting the RF sweep frequency higher and lower than the clock frequency and watching for the point that the loop frequency changed as the deceptive signal was increased in power. The maximum limits of loop pull off were equal to the capture range of 40% of the clock frequency. The power required for loop capture was on the order of that required to cause zero intelligibility in the noise and CW interference cases. The swept audio technique used a little less power for loop capture, probably because it looked more CW like than the swept RF.

The second technique used a carrier at 21.4MHz and an amplitude modulated swept audio that varied $\pm 40\%$ of the clock frequency. The power was slowly increased until the loop lock was broken. Again, the signal leaked through the correlator SAW device and overcame the signal from the first SAW device to capture the loop.

Table 5 indicates the SAW power budget used for determination of the signal strength in these tests. SAW device insertion loss was measured at 43.3dB rather than the 66dB listed in Ref. 10.

CHANNEL A		CHANNEL B	
INPUT	-13.9 dBm	INPUT	-13.0 dBm
INS. LOSS	-43.3 dB	INS. LOSS	-43.3 dB
OUTPUT	-57.2 dBm	OUTPUT	-56.3 dBm
CHAN. AMP	+30.0 dB	CHAN. AMP	+30.0 dB
OUTPUT	-27.2 dBm	OUTPUT	-26.3 dBm

ALGEBRAICALLY SUMMED OUTPUT
FROM SAW DEVICE #2

	-67.5 dBm
AMPLIFIER	+30.0 dB
INPUT TO PEAK DETECTOR	-37.5 dBm

f_{clock} = 170KHz

Rf pulse width = .98Us

TABLE 5

SURFACE WAVE DEVICE POWER BUDGET

V. CONCLUSIONS

The effort presented in the previous chapters resulted in a laboratory model that satisfied the original premise of building a system that would cost a potential adversary more than the information carried might be worth to routinely intercept and demodulate it. This cost can be made higher by permuting the converter pins with a permuter board and routinely changing the permutations daily. The cost can be made prohibitive if an encryption device such as used in the VINCENT system were included. With the advent of such encryption systems and multipurpose radio sets, that idea might have significant importance if a set of the type designed here were to have a chance at production.

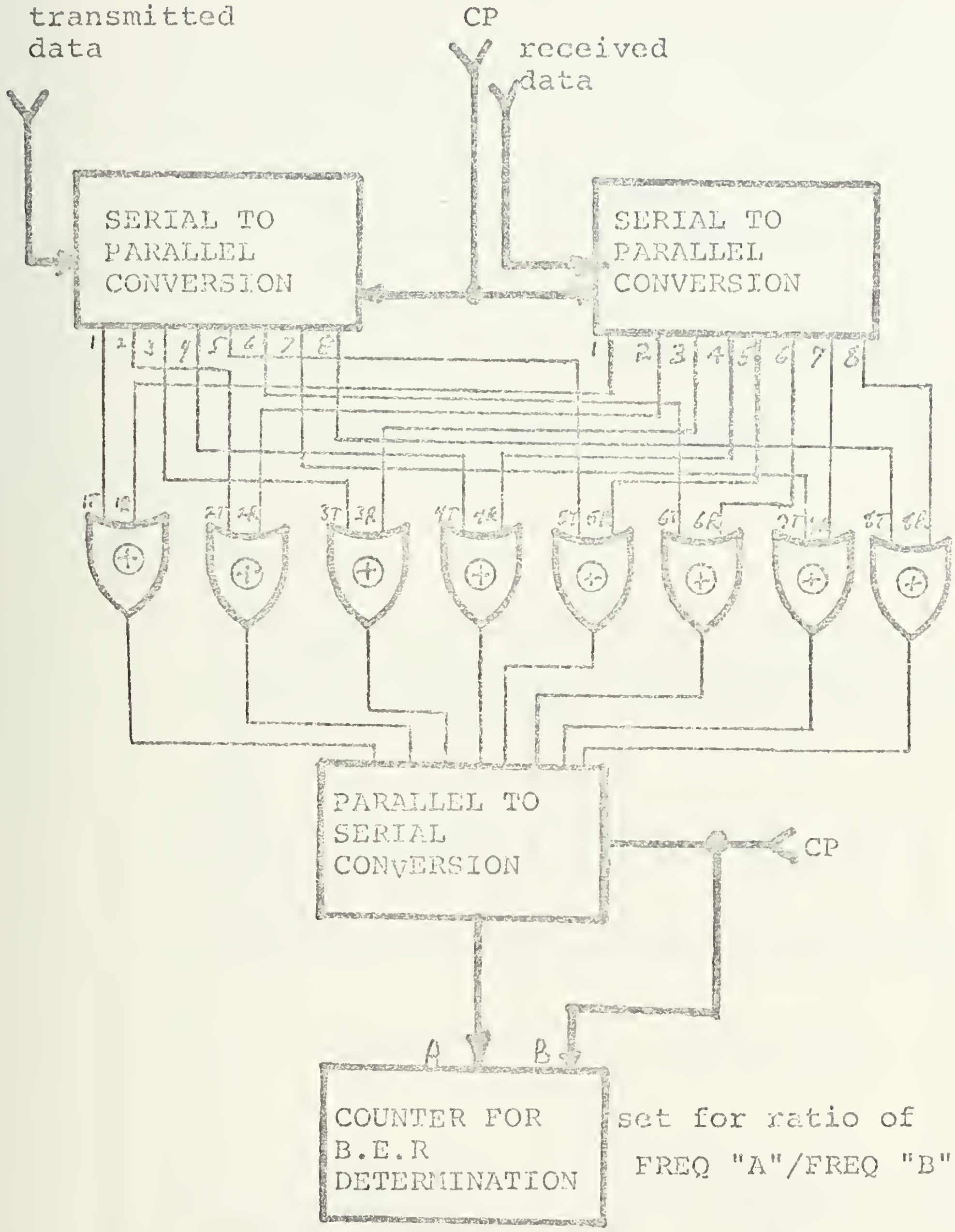
The low low power multichip CCD devices could provide a longer data frame and more choice as to where in the frame the data is to be placed. The entire A/D converter, P/N code generator and pulse stuffing scheme could easily be put "on chip." With sufficient ingenuity, other housekeeping functions might also be placed "on chip." Should a potential manufacturer not choose to go to CCD, significant power savings could be realized by redesigning the CMOS chips.

The tests that were performed were those minimum engineering tests required to assure performance. Other tests are certainly possible and should be designed to probe the system limitations and weaknesses.

A suggested circuit for bit error rate measurement is shown in Fig. 43. The counter can either count directly or measure a ratio as shown. When no errors occur, the output of the parallel to serial converter will be all zeros. An error will cause an exclusive OR output and cause an output from the converter to the counter.

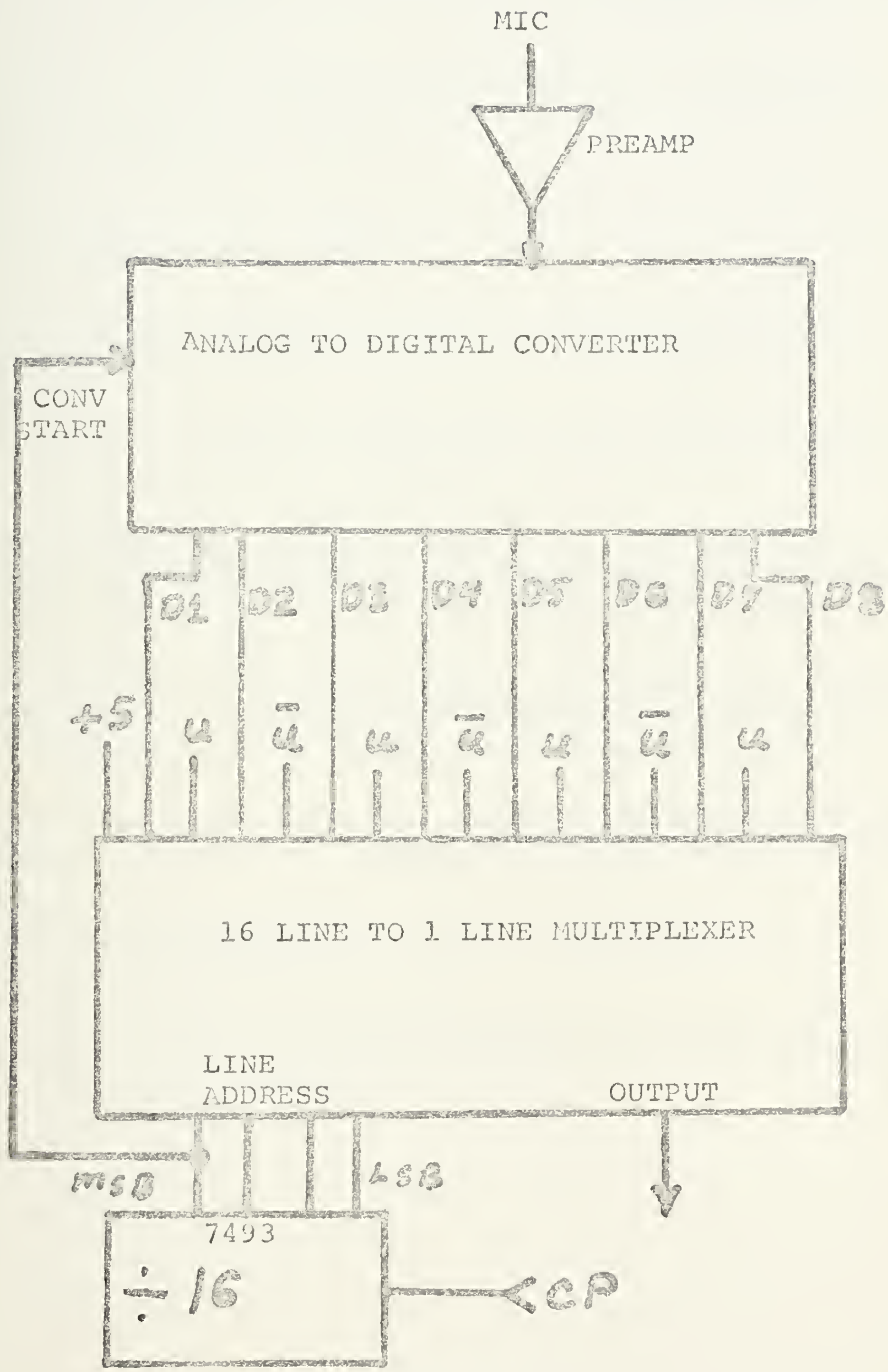
Many alternate designs for this thesis project are possible. An SPU of simpler design and lower power requirements is shown in Fig. 44. The design was bread boarded and tested. There was no discernable difference in the outputs of this device and the one built in the box.

Much remains to be done with this project. First a redesign for smaller space and power power consumption needs to be accomplished. Secondly, a design must be made for a frequency converter and an RFPA. The receiver front end and RFPA must both employ some form of non phase distorting amplification. Even the antenna coupling design could be a problem as no ringing or phase distortion can be allowed there either. New SAW devices are available which will bring the circuit size down. They may have matching problems however. Digital generation of the required sequences and RF pulse generation using these sequences without SAW devices should also be investigated and compared with the SAW device results. Reduction in number of power supply voltages required should be accomplished also in order that more cost effective single voltage batteries be used.



SUGGESTED SETUP FOR BIT ERROR RATE (BER) DETERMINATION

FIG. 43



SEND PROCESS UNIT ALTERNATE DESIGN
FIG. 44

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